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THE PHILIPS INDUSTRIES

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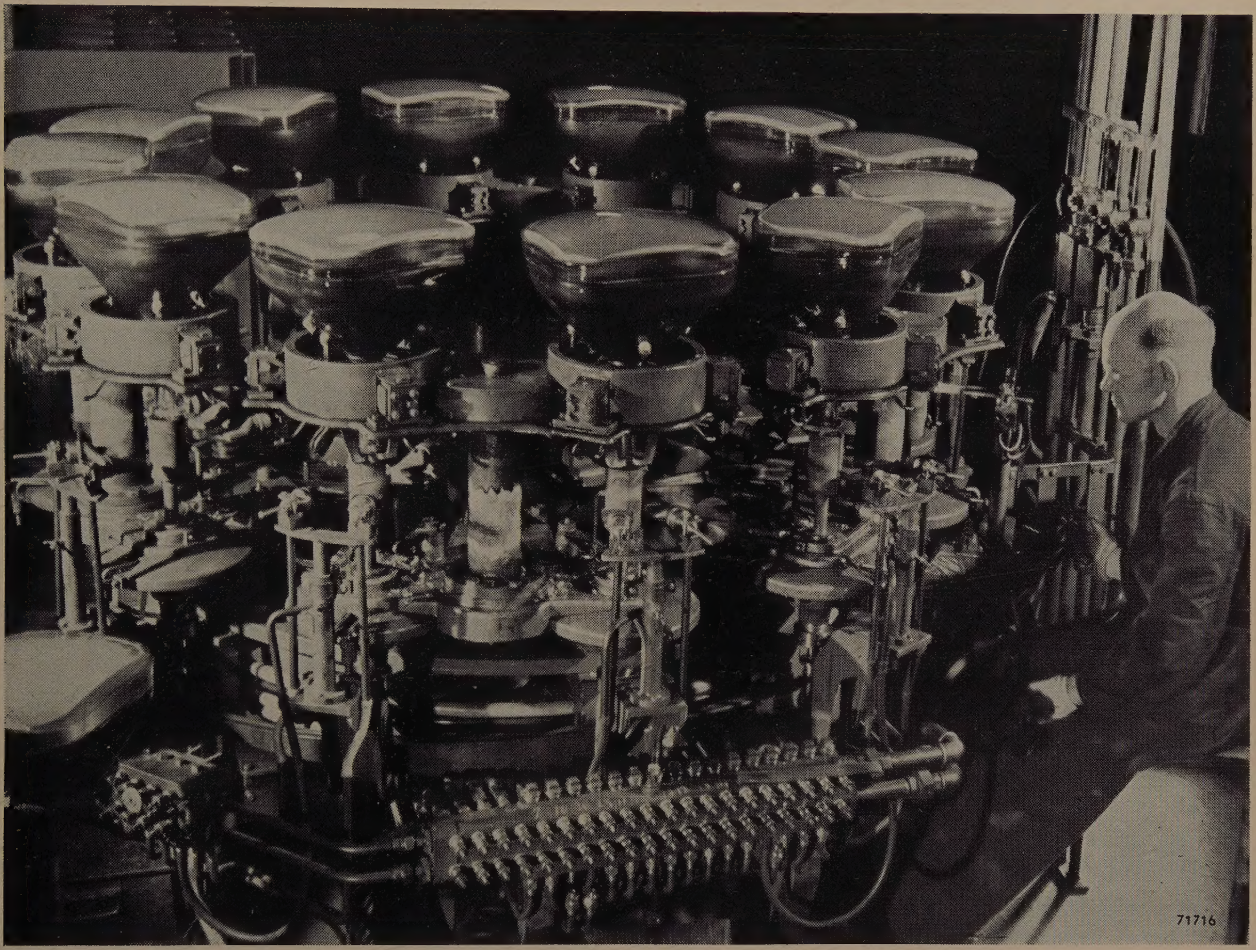


Photograph Walter Nürnberg

ELECTRONIC TUBES

A synopsis by J. L. H. JONKER. 621.385

The following article covers the main points of an address by Prof. Dr. J. L. H. Jonker on his formal acceptance of the office of professor at the Technical University at Delft on 19th March 1952. Although the extent of his subject did not admit of more than a brief review, this "bird's-eye view", which brings many widely separated aspects into a common focus, has an appeal of its own.

Nothing but the greatest admiration can be felt by anyone tracing the rapid development of electronics, that is, electronic tubes and their applications, which has sprung from the phenomenal growth of radio in the first half of the twentieth century.

It was roughly at the turn of the century that Hertz, Sir Oliver Lodge and Marconi carried out the experiments that led to the successful introduction of telegraphy without wires by means of high-frequency oscillations.

In the short span of a few decades radio has

undergone almost unbelievable technical development and improvement, and it has achieved a revolution in the interchange of ideas between individuals and peoples alike. It has come to assume such importance in the lives of almost everyone that throughout the whole evolution of mankind there have been few technical developments to equal it.

The early days of radio

This development of radio, which took the world by storm, founds its origin in the experiments already mentioned. Channels were established across the oceans, first telegraphically and later by telephone, up to the point where radio contact now covers the whole globe.

In the early years, technical knowledge and literature on the subject were in the nature of things very limited. Because the individual experimenter was, however, in very many instances able to achieve valuable results and improvements using only simple and often home-made equipment, many felt themselves drawn towards this sphere and this is undoubtedly one of the causes of the very rapid advances made. Those who were privileged to witness at close quarters some part of this inspiring early period may look back on those exciting days with some regret, comparing them with conditions as they are today, now that improvement and research into specific problems are in the hands of a corps of specialists having the most ingenious laboratory equipment at their disposal.

As a result of the work of these specialists, literature on the subject of radio would now constitute a whole library in itself, whereas, on the other hand, the ceaseless flow of publications presents to the expert the problem of keeping in touch with everything that is of interest in his work. Here, too, a too copious emission results in so great a space charge that the ultimate object is defeated. In the United States this urgent problem has recently given rise to an investigation into the statistical distribution of such publications among the various periodicals; the result of this investigation has been reported — in another publication! — in the periodical containing the largest number of such publications [1]).

It is because of the initial circumstances already outlined that history can point to one specific invention of one man, as being as it were the lever which released radio from its trammels. This was the start of the subsequent fantastic growth and was, in the words of the Nobel-prize winner Rabi: "so outstanding in its consequences that it almost ranks with the greatest inventions of all time".

This invention is the *triode* of Lee de Forest.

The advent of the triode

In seeking for a better detector of radio signals Lee de Forest [2]) in 1907 introduced a grid between the cathode and anode of the existing diode devised by Sir John Ambrose Fleming, thus obtaining a means of controlling the flow of electrons between

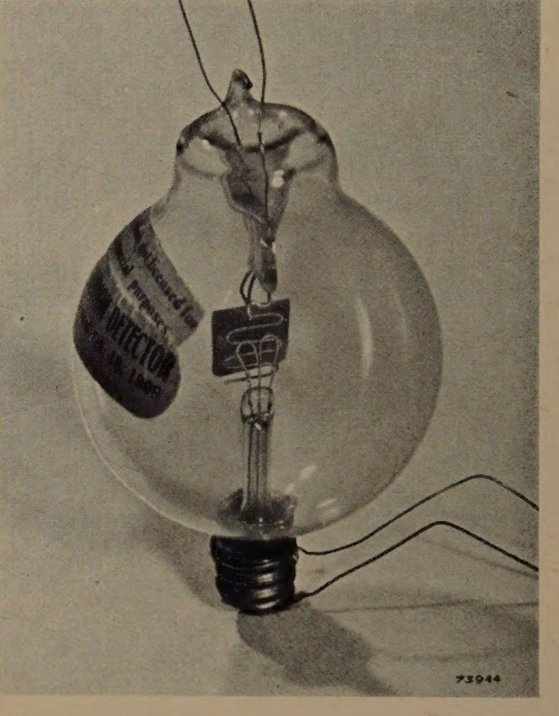


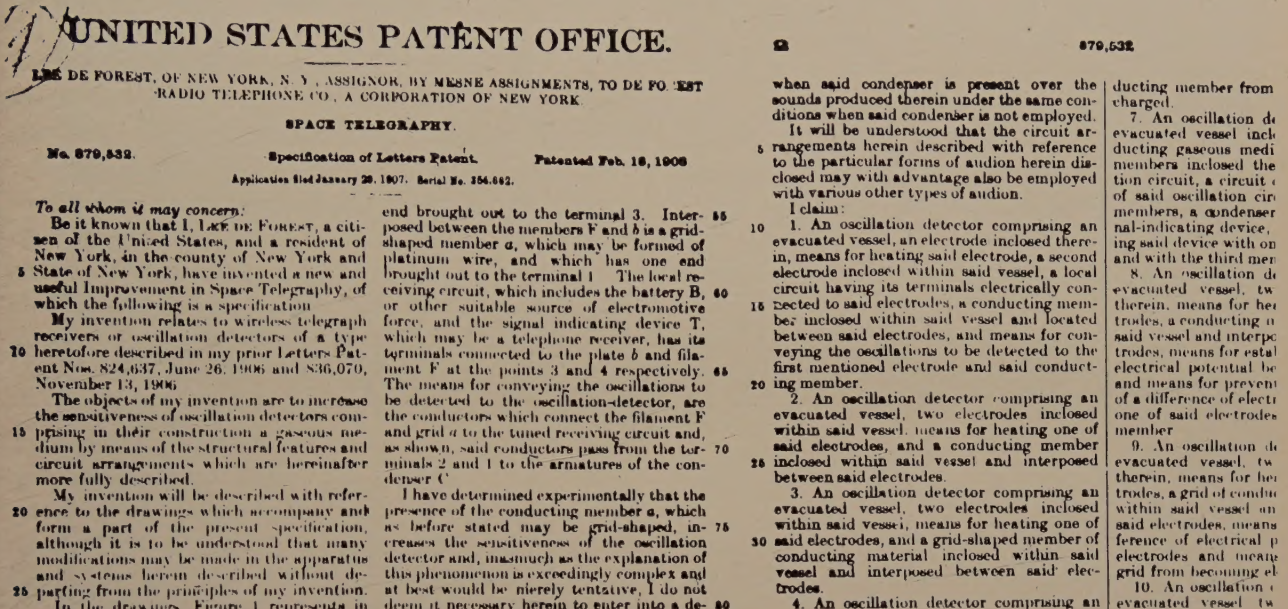
Fig. 1. The original triode ("audion") of Lee de Forest.

cathode and anode without the consumption of any energy (*figs 1* and *2*). Years elapsed before the operation and possibilities of this new invention were fully appreciated, in consequence of which De Forest became so short of funds that he had to let his European patents lapse! In 1911 Robert von Lieben gave a demonstration of the amplification properties of the triode at the Berliner Physikalische Gesellschaft. In 1913 different research workers simultaneaously invented the "feed-back", which permitted the triode to be used as a very effective generator of high-frequency oscillations. The last-mentioned invention gave rise to the longest case of patents litigation in this field in history [3]), a contest which was not settled until 1934, in favour of De Forest.

The invention of the triode, which marked the commencement of the electronic era, gave to radio

at once a better detector, the long sought amplifier and an oscillator. These various possibilities provided the stimulus for a careful study of these tubes and for such modifications as would ensure optimum results in any function in transmitting and receiving equipment, and this has in turn produced the innumerable varieties of electronic tubes which we know today.

reproduction and the oscilloscope. In the years that followed, research penetrated the realms of ultra-short waves, atmospheric interference, frequency modulation and television, apart from innumerable industrial applications of electronics. In each case the electronic tube was the key that opened the door to the new sphere, and often, in new forms, to entirely new possibilities. This



UNITED STATES PATENT OFFICE.

LEE DE FOREST, OF NEW YORK, N. Y., ASSIGNOR, BY MESNE ASSIGNMENTS, TO DE FOREST RADIO TELEPHONE CO., A CORPORATION OF NEW YORK.

SPACE TELEGRAPHY.

No. 879,532. Specification of Letters Patent. Patented Feb. 18, 1908.

Application filed January 29, 1907. Serial No. 354,662.

To all whom it may concern:

Be it known that I, LEE DE FOREST, a citizen of the United States, and a resident of New York, in the county of New York and State of New York, have invented a new and useful Improvement in Space Telegraphy, of which the following is a specification.

My invention relates to wireless telegraph receivers or oscillation detectors of a type heretofore described in my prior Letters Patent Nos. 824,637, June 26, 1906 and 836,070, November 13, 1906.

The objects of my invention are to increase the sensitiveness of oscillation detectors comprising in their construction a gaseous medium by means of the structural features and circuit arrangements which are hereinafter more fully described.

My invention will be described with reference to the drawings which accompany and form a part of the present specification, although it is to be understood that many modifications may be made in the apparatus and systems herein described without departing from the principles of my invention.

In the drawings, Figure 1 represents in

end brought out to the terminal 3. Interposed between the members F and *b* is a grid-shaped member *a*, which may be formed of platinum wire, and which has one end brought out to the terminal 1. The local receiving circuit, which includes the battery B, or other suitable source of electromotive force, and the signal indicating device T, which may be a telephone receiver, has its terminals connected to the plate *b* and filament F at the points 3 and 4 respectively. The means for conveying the oscillations to be detected to the oscillation-detector, are the conductors which connect the filament F and grid *a* to the tuned receiving circuit and, as shown, said conductors pass from the terminals 2 and 1 to the armatures of the condenser C.

I have determined experimentally that the presence of the conducting member *a*, which as before stated may be grid-shaped, increases the sensitiveness of the oscillation detector and, inasmuch as the explanation of this phenomenon is exceedingly complex and at best would be merely tentative, I do not deem it necessary herein to enter into a de-

2 879,532

when said condenser is present over the sounds produced therein under the same conditions when said condenser is not employed.

It will be understood that the circuit arrangements herein described with reference to the particular forms of audion herein disclosed may with advantage also be employed with various other types of audion.

I claim:

1. An oscillation detector comprising an evacuated vessel, an electrode inclosed therein, means for heating said electrode, a second electrode inclosed within said vessel, a local circuit having its terminals electrically connected to said electrodes, a conducting member inclosed within said vessel and located between said electrodes, and means for conveying the oscillations to be detected to the first mentioned electrode and said conducting member.

2. An oscillation detector comprising an evacuated vessel, two electrodes inclosed within said vessel, means for heating one of said electrodes, and a conducting member inclosed within said vessel and interposed between said electrodes.

3. An oscillation detector comprising an evacuated vessel, two electrodes inclosed within said vessel, means for heating one of said electrodes, and a grid-shaped member of conducting material inclosed within said vessel and interposed between said electrodes.

4. An oscillation detector comprising an

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Fig. 2. Extracts from the United States patent granted to Lee de Forest, covering the invention of his triode. The "grid-shaped member" is mentioned in the third claim.

Rapid development

During and after the first world war we have seen the efforts that were directed towards producing electronic tubes of high and very high power for the establishment of world-wide radio communication. Then came a remarkable offshoot in the field of electronics: the radio broadcast, commenced with noble idealism, but later to be misused for commercial advertising and political propaganda. Radio broadcasting has in many parts of the world called into existence enormous electronic industries whose annual output is now estimated at some 500 million tubes and a good 30 million receivers, the turnover for 1952 being estimated at many gigadollars. In this way a new sphere of employment has been created for hundreds of thousands in the manufacture and application of electronic tubes.

The young but rapidly expanding industries with their many specialists soon found more scope in other directions, to mention only sound amplification systems, soundfilm, improved gramophone

applies equally to the latest branches of electronics, developed during the last war and subsequently perfected, e.g. radar, sonar and other aids to navigation, radio relay stations and computing equipment.

Electronics in general is much indebted to the second world war, that black page in the history of mankind, since technical development was urged forward by the exigencies of the war without being hampered by economic and commercial factors. Many of the results obtained in this way have since proved to be of great scientific, technical and commercial interest and have been adopted or further developed for peaceful occupations; moreover, new techniques are continually being added to the already wide field of electronics, such being for example radio astronomy, colour television and microwave spectroscopy. The very extensive scope of application of electronic tubes and the consequent diversity of their form may be illustrated with reference to some of the extremes among the dimensions involved.

The diversity of present-day electronic tubes

It is possible by means of electrometer triodes to measure currents of 10^{-15} amperes, whilst in transmitting tubes and gas-filled rectifiers, current peaks of 100 amperes or more occur. Using electronic tubes we can measure alternating voltages of 10^{-6} volts (it should be recognized here that in that case out of a possible 10^{18} available electrons in an amplifying tube, only about 10^{10} are controlled). On the other hand, an installation is being built by the Stanford University in California (a "linear accelerator") which, with the aid of enormous transmitting tubes (klystrons), operated at 300 000 V, will accelerate electrons to the equivalent of a thousand million volts [4]). Electronic tubes can be made to operate within a range of from 0 to 10^{11} cycles per second. The weights of conventional types of tube may differ one from the other by a factor of as much as 100 000.

Within these extreme limits there is an enormous variety of design among electronic tubes; a cautious estimate places the number of different commercial types at about 20 000.

It is not within our scope to give even the briefest review of all these different designs. We shall confine ourselves to a short discussion of some of the more important applications only.

Receiving tubes

The most important field of application of tubes numerically is without a doubt to be found in radio receivers. In the last decades there has been a refinement in the design of the receiving tube, consisting in the attainment of smaller electrodes and closer spacing between the electrodes, that have resulted in an over-all reduction in the size of the tube (*fig. 3*). The inherent advantages are, amongst others, lower current consumption and better short-wave characteristics; manufacturing costs have dropped owing to more rapid production, lower consumption of materials, smaller stores and reduced transport costs. The receivers in which the tubes are used could be made more compact in consequence.

Intensive research has taken place into the many physical phenomena occurring in electronic tubes that may be useful or detrimental, such as thermal, secondary and photo-emission (*fig. 4*), fluorescence, gas discharge, space charge, potential fields, electron paths, noise and, in short-wave operation, transit-time effects and damping. Investigations into these effects have not by any means ended; on the contrary, efforts are still being made, based on a growing knowledge of these often highly complex phenomena, to improve the performance of the tubes. It was, for example, possible to improve the efficiency of pentodes considerabely by taking into account in the design the paths of the electrons through the potential fields of the grids [5]). In most cases, however, a computation of these potential fields and the paths of the electrons within them would take up far too much time, or would prove to be impossible or possible only under very restricted conditions. By means of models of the electrode system in the "electrolytic tank" or on a rubber sheet (*figs 5* and *6*), a better idea is obtained

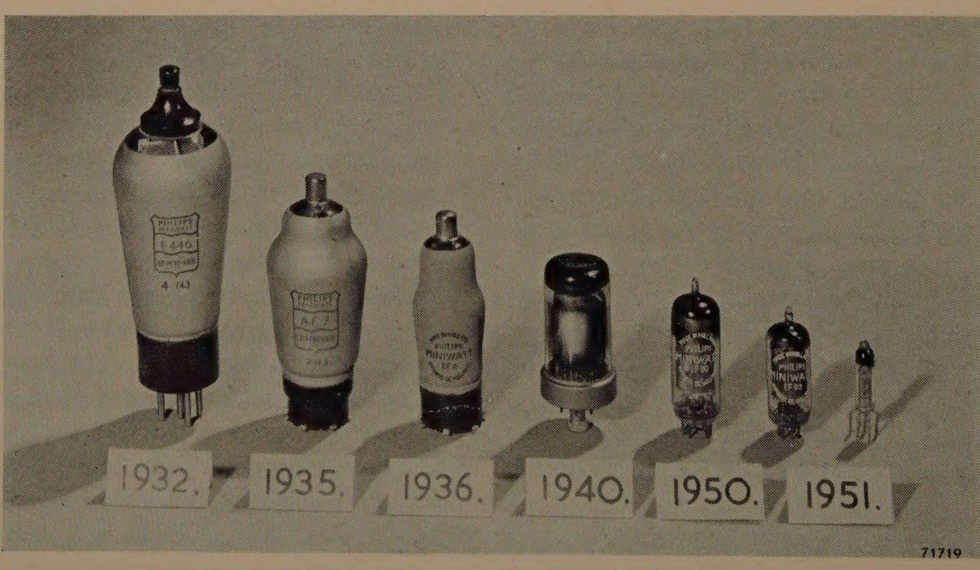
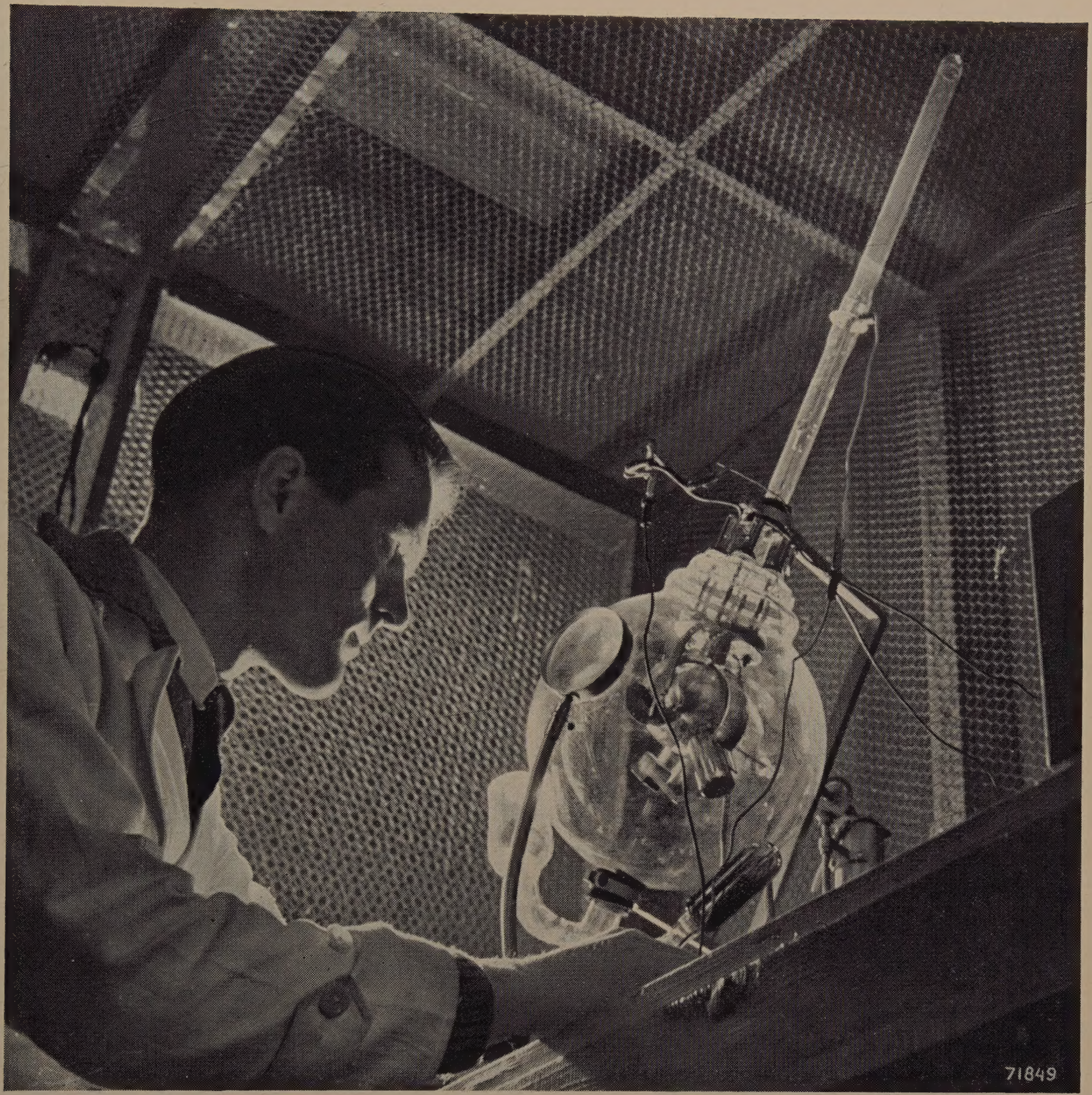


Fig. 3. In the course of time the dimensions of radio receiving tubes have been reduced again and again. All the tubes shown are pentodes.



Photograph Walter Nürnberg

Fig. 4. Apparatus employed for the measurement of secondary emission of solids. For a description see J. L. H. Jonker, The angular distribution of the secondary electrons of nickel, Philips Res. Rep. **6**, 372-387, 1951 (No. 5).

of what occurs inside the tube and thus of the means of simplifying the computations [6]).

In the course of time endeavours to secure improved tube performance in the receiver have on many occasions shown the desirability of including more than one grid between cathode and anode, and in this way the well-known tetrodes, pentodes, hexodes, pentagrids, octodes, having respectively 2, 3, 4, 5 and 6 grids, originated. The record number of grids so far is found in the enneode (7 grids) employed as a phase detector for F.M. signals [7]). Small dimensions and a high degree of accuracy in the arrangement of the electrodes and their leading-in wires ensure that these broadcast tubes will operate effectively at wavelengths down to a few metres [8]). The special form of construction necessary for tubes intended to work at still shorter wavelengths will be considered under a separate heading.

Whereas the configurations of the electrodes in tubes for the various functions in receivers have become more or less stabilized in recent years, the same cannot be said of tubes for television reception. The short wavelengths, the wide frequency band and the sav-tooth voltages required for the motion of the cathode ray over the screen impose other requirements on the tubes than those necessary for ordinary broadcast reception. These

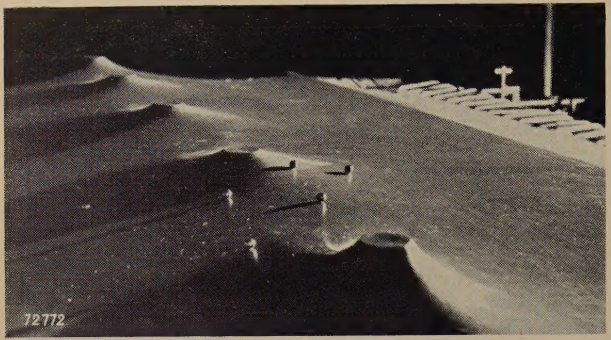


Fig. 5. This deformed rubber sheet reproduces the potential field in the cross section of a plane triode having 5 grid rods, for given values of the grid and anode potentials. Steel balls rolling over the sheet describe the same paths as electrons in the potential field.

requirements are again best met by tubes which are specially designed for the purpose [9]). A similiar situation will arise in the future in the case of colour-television receivers, and we have every justification for anticipating that new designs will be evolved which will render the receivers more sensitive and less complicated.

Efforts to reduce the number of tubes per receiver by introducing special types would at first sight appear to constitute a laborious, suicidal policy on the part of the manufacturer. This is by no means the case, for simpler and less expensive receivers can be sold in larger quantities. Apart from that, it is difficult to keep in check improvements and simplifications of techniques.

Transmitting tubes

In transmitting tubes, the main function of which is to supply considerable power with a high degree of efficiency, the maximum temperature which the tube materials are able to sustain determines the limit of utilization and, if such be possible, technology here plays an even more important part than in receiving tubes. Amongst classical types, i.e. the triode, tetrode and pentode, the volume of the tube for a given output power has in recent years been reduced by a factor of from 5 to 10 (*fig. 7*) by employing high-melting-point materials with good radiation properties for the electrodes, non-volatile getters and new types of glass having improved thermal and electrical characteristics, as well as by making a special study of the dimensioning of the electrodes and their leading-in wires [10]). The reduction in inter-electrode capacitances brought about by thus curtailing the size of the tubes is particularly important at the very short wavelengths, where low capacitance is a *conditio sine qua non*. The maximum permissible loading of such small electrodes, however, sets a limit to the output power. When it is not possible with given electrode dimensions to dissipate enough of the heat by radiation, forced cooling with air or water is employed (*fig. 8*), for which purpose highly efficient cooling systems have been designed in which the most advantageous turbulence of the coolant is ensured [11]). In the future, especially among short-wave transmitting tubes, further

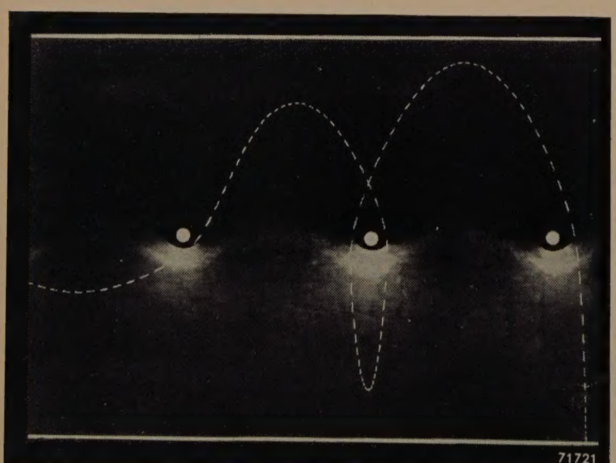


Fig. 6. Example of path of ball on the rubber sheet as photographed in intermittent light. The photograph shows that under certain conditions the electron can undergo such deflection between the grid wires that it is flung beyond the space between the electrodes.

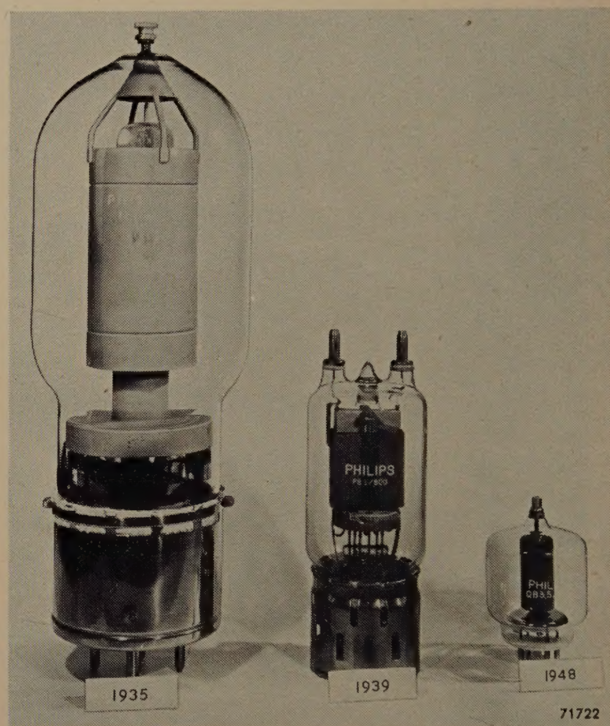


Fig. 7. Evolution of the transmitting tube with the years. All three types deliver about 1000 W; the latest of these is 14,5 cm in height.

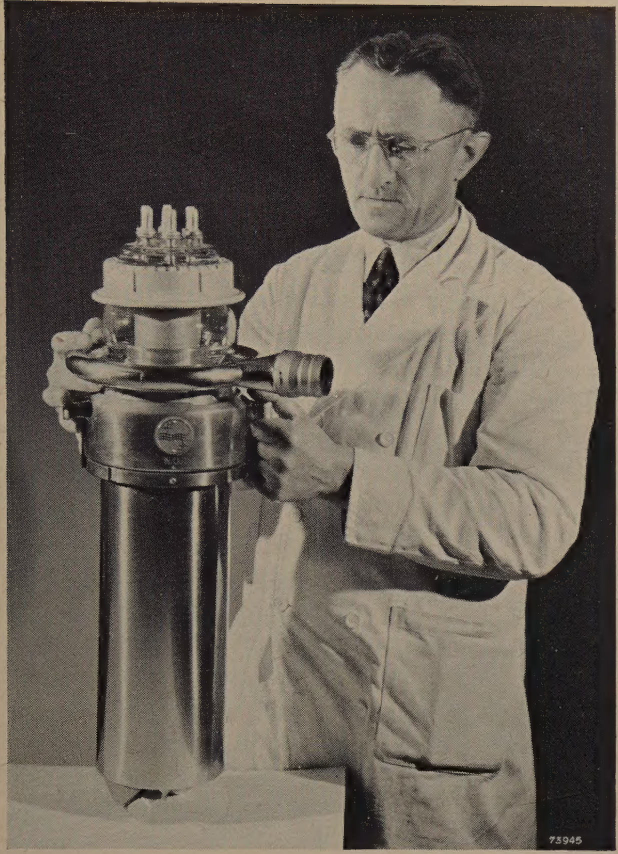


Fig. 8. Water-cooled transmitting tube for 100 kW output, with its anode water-cooling jacket.

improvements in consequence of continued technological research, as also further refinements in the electrode system, are to be anticipated.

Tubes for ultra-short waves

The urge towards the new and unknown has prompted many experimenters to investigate shorter and still shorter wavelengths. For the original radio links over great distances the long-wave range of 1000-20 000 metres was employed, but techniques have in the course of time tended more and more towards shorter wavelengths for such purposes. As to this, we are much indebted to amateurs all over the world for the part they played in this pioneer work.

At present normal communication transmitters, amongst which those intended for broadcasting, operate at wavelengths ranging from 25 kilometres to about 1 metre, whilst for normal television wavelengths of several metres and for colour television decimetric waves are used. Centimetric waves are now used for radio relay stations and for radar; millimetric waves serve physical science in the study of the characteristics of molecules and atomic nuclei by means of the absorption spectroscopy of gases [12]).

The ordinary receiving tube is not effective at decimetric and shorter waves, when the frequency is increased to such an extent, the gain per stage drops below unity. The causes of this, viz. electrical losses, radiation and transit-time effects, are in part ascribable to the characteristics of the circuit at such wavelengths; in part, too, they are of electronic origin [13]). Here, other forms of tuning devices, as well as tubes of different design are needed to ensure success; for example, in the centimetric and millimetric wavelengths, the circuit consists of transmission lines, coaxial cables, wave-guides and cavity or rod-shaped resonators [14]), the radiation and I^2R losses of which can be made appreciably lower than those of the conventional types of coil employed for the longer wavelengths. It has been found possible to extend the working range of the tubes towards the shorter wavelengths, firstly by reducing the size and spacing of the electrodes so as to cut down capacitances and transit-time effects and, secondly, by making the leading-in wires for the electrodes function as part of the resonant circuits. The latter step has led to the use of disc-shaped or annular electrode connections in short-wave tubes; the metal discs, being thus part of the resonant circuit, are sealed laterally through the wall of the glass envelope, so that the part inside can serve as support for an electrode (*fig. 9*).

As to the reduction in the electrode spacing, it has in some cases been found possible to cut down the space between the control grid and cathode to

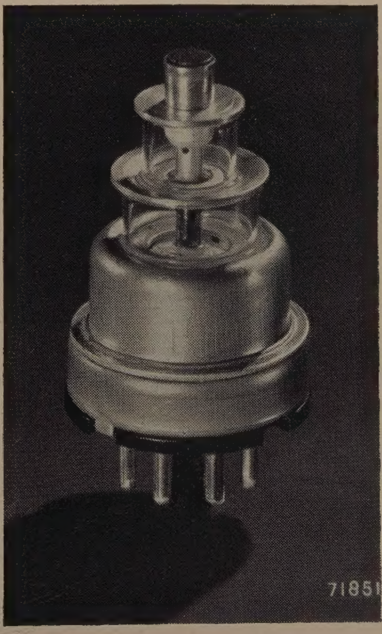


Fig. 9. "Disc-seal triode" type EC 56. Grid and anode are carried by metal discs sealed laterally into the glass wall; the outer part of the discs forms part of the resonant circuit.

as little as a few hundredths of a millimetre [15]). So that the controlling action shall not be lost in consequence of this, the mesh of the grid has to be unusually fine; to this end taut tungsten wires 10 microns in thickness are used, this being many times thinner than human hair (*fig. 10*). Transit times in an oscillator can be reduced still further by the use of higher voltages, which, in order not to overload the electrodes, are applied in pulses; the electrodes thus have time to cool between two pulses.

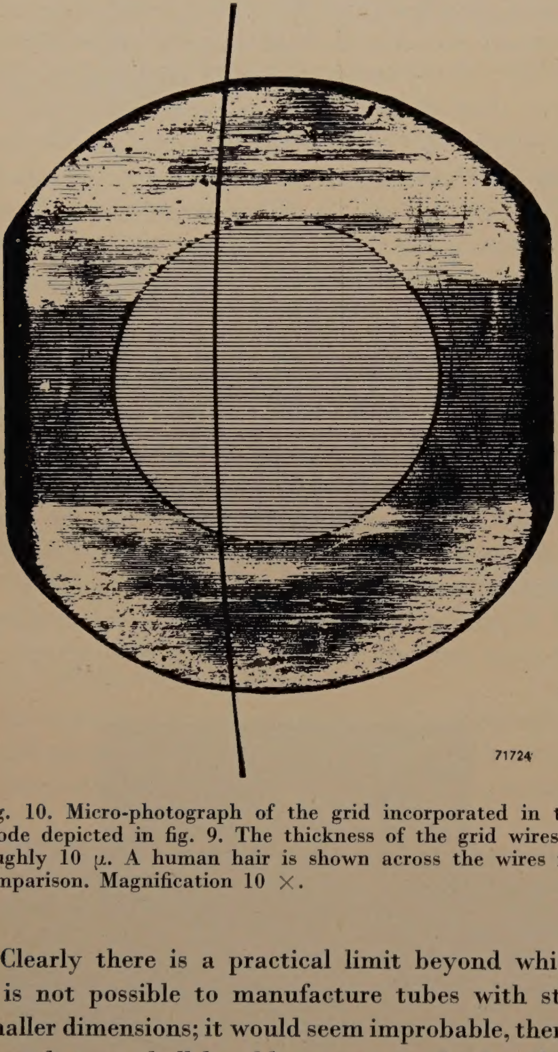


Fig. 10. Micro-photograph of the grid incorporated in the triode depicted in fig. 9. The thickness of the grid wires is roughly 10 μ. A human hair is shown across the wires for comparison. Magnification 10 ×.

Clearly there is a practical limit beyond which it is not possible to manufacture tubes with still smaller dimensions; it would seem improbable, therefore, that we shall be able to get any further than wavelengths of some centimetres by this means. Ultimately the limited loading capacity of the electrodes, as well as the impracticability to fix electrodes operating at elevated temperatures with extremely small spacings, will prove the deciding factors. Another limitation is found in the size of the cathode, since emission, and therefore also the output power, decrease with the area of the cathode. Increased emission through higher specific loading can be achieved only at the expense of the life. Under pulsed conditions, however, the specific emission of the ordinary oxide-coated cathode can be considerably increased [16]).

In order to lengthen the life when the specific emission is increased, special cathodes have been developed, such as the L cathode in which the same alkaline earth oxides are used as for the conventional oxide-coated cathode, but applied behind a porous layer of sintered tungsten [17]).

If high power is required at very short wavelengths, it is necessary to resort to electrode systems whose dimensions are of the same order as the wavelengths at which these systems should operate; this, now, is the case in tubes whose action is based on a utilization of transit-time effects, as in magnetrons which during the second world war played such an important part in radar for locating submarines and aircraft, as also in klystrons, travelling-wave tubes and similiar systems [18]). In such tubes the transit time of the electrons is of the same order as the periodic time of the alternating current; thus the electrons are able to impart energy to the high-frequency field which is thereby increased so that oscillations are produced. Although such systems do not excel by reason of their efficiency, they have the advantage that they are large enough with respect to the short wavelength not only to dissipate the heat generated, but to be manufactured without serious mechanical difficulties.

For generating waves of 3 and 10 cm, klystrons and magnetrons have been made that supply a peak output of respectively 10 and 5 megawatts, with efficiencies of 30% and 60%. On the other hand, a 3 cm magnetron has been manufactured for a peak output of 100 W which weighs not more than 1 kg including the permanent magnet (*fig. 11*).

The shortest wavelength for which magnetrons can be manufactured is in the region of a few millimetres, the efficiency being rather low. It has also proved possible to obtain a small amount of power, as higher harmonic of the fundamental wave, at a wavelength of roughly 1 millimetre [19]).

Klystrons have been made for a continuous oscillation at a wavelength of a few millimetres, but, as the circuit losses are inversely proportional to the square root of the wavelength, the efficiency at such wavelengths is very low.

These tubes are accordingly not suitable for amplification purposes. The travelling-wave tube is much less susceptible to poor circuital properties in this range of frequencies, as it operates without a resonant circuit.

In these tubes a beam of electrons, usually concentrated by means of a magnetic field, is shot in axial direction through an elongated wire helix,

the signal being applied to the cathode side of the helix. When the speed of propagation of the beam and that of the field of the travelling wave in the axial direction are made almost identical, the signal is amplified along the length of the helix, at the expense of the beam. With greater losses in the helix it is only necessary to employ a longer helix for the same amount of gain.

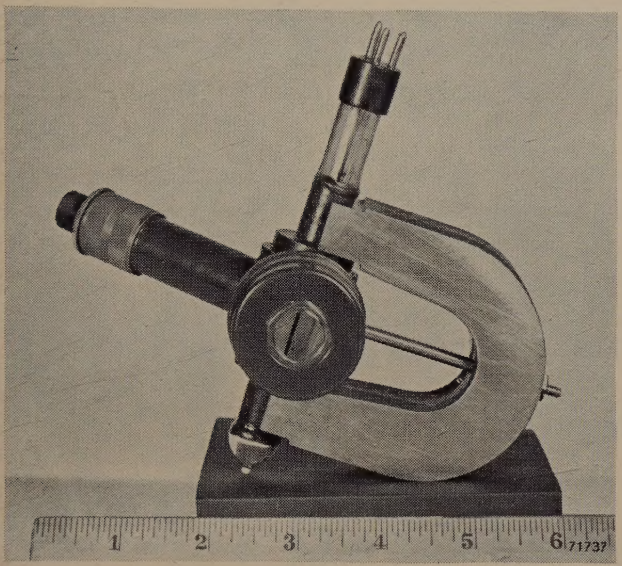


Fig. 11. Magnetron for a wavelength of 3 cm designed by Philips Laboratories, Irvington, N.Y., U.S.A. This magnetron is not intended for radar, but for a radio beacon. Peak output 100 W at an anode voltage of 800 V, heater power 2 W. Total weight including the permanent magnet: approx. 1 kg. The size may be seen from the inch rule placed in front of the magnetron. (Photo by courtesy G. A. Espersen and B. Arfin, Tele-Tech. **10**, No. 6, p. 50 and No. 7, p. 30, 1951.)

On this principle tubes have been made that will provide ample gain at a wavelength of only a few centimetres, and the success achieved with these tubes on such wavelengths has stimulated experiments with numerous similar systems, in which the helix is replaced by rods, wires, a wave-guide, or a second electron beam.

There is still an enormous amount of scope for microwave experimentation, if only by reason of the fact that among the many different methods of making a high-frequency alternating field and a constant electric and magnetic field cooperate spacially with a beam of electrons, only few have been investigated in tubes. More systematic research than has hitherto taken place might yield remarkable possibilities.

In many laboratories all over the world efforts are being made along widely divergent lines to penetrate further into the interesting problem of generating and amplyfing microwaves.

Cathode-ray tubes

In cathode-ray tubes use is made of a narrow beam of high-speed electrons to produce a luminous image. This is achieved by deflecting the beam, or cathode ray, electrically, so that it travels over a surface coated with substances that emit light in consequence of the electron bombardment. Originally these tubes were employed in the cathode-ray oscilloscope. The coming of television has imposed new conditions on the quality, and mass production, too, has brought other considerations to bear. A distinction is made between two systems in television, viz. direct and indirect vision of the images. Until now the former has been the more widely used of the two systems, but this has led to the making of tubes with larger and larger screens, seeing that small images tend to tire the viewer and to limit the number of viewers per receiver. In order to reduce the weight of such large tubes, the heavy glass cone has been replaced by a lighter one of metal [20]) and the inconveniently large dimensions have been cut down by introducing a rectangular screen to take the place of the round one (see also frontispiece), and shorter electrode systems for generating and deflecting the electron beam. It has also been found possible to shorten the cone by employing a narrower electron beam which is capable of deflection through larger angles (90°) (*fig. 12*).

It may well be asked whether it is desirable to go any further in this direction to produce still larger pictures. The presence in the living room of a vacuum tube 75 cm in diameter, as recently produced in the United States, would seem to be anything but desirable, even though the danger of implosion due to the atmospheric pressure of roughly 4 tons on the glass screen need not be considered high in a properly designed tube. In order to withstand such forces, the screen has to be very thick, and also more or less convex, which means a certain amount of distortion for the viewer. Lastly it may be said that a tube of such dimensions presents quite a problem for the set manufacturer in the design of a cabinet that will be suitable for installing in the living room.

If large or very large pictures are demanded, the indirect system is much to be preferred; a cathode-ray tube of quite small dimensions but very high luminous intensity then provides an image that can be projected on to a flat screen by means of lenses or a concave mirror [21]). The high luminous intensity of the screen required in this case imposes very stringent conditions on the luminescent substances from the point of view of loading.

Camera tubes for television

Developments in camera tubes for television have led to magnificent and most interesting results. We have been given the iconoscope, the image iconoscope (*fig. 13*), the orthicon and the image orthicon, associated with such names as Zworykin, Rose and Iams [22]). These tubes, which convert the optical image into an electrical signal and which rank amongst the most complex of electronic inventions, were the means of appreciably hastening forward the realization of practical television. The technological problems to be solved were great. For instance in the iconoscope the optical image is translated into a pattern of electrical charges by means of a screen or mosaic about 100 sq. cm containing several milliards of minute photo-cells, each insulated from its neighbours, in a density of roughly 360 000 per sq. mm. For comparative purposes it may be pointed out that the concentration of the rods and cones in the retina of the human eye is at most 20 000 per sq. mm, which means that the artificial retina of the iconoscope improves on nature to the extent of almost 20 times (*fig. 14*). In order that the image may be transmitted, it is scanned in lines at a speed of thousands of metres per second by a beam of electrons, the necessary signal being obtained by "measuring" the charges of the photo-cells along those lines.

Means have been found of ensuring a very high sensitivity to light in the image orthicon, namely 10 times higher than that of the fastest photographic material; hence a serviceable image is obtained by the light of a candle, or by moonlight. This has been achieved, firstly, by using for the scanning a beam of electrons having a comparatively low velocity, thus eliminating the interfering secondary emission from the mosaic inherent in the iconoscope and, secondly, by making effective use of secondary emission by including in this already complicated tube a multi-stage electron multiplier.

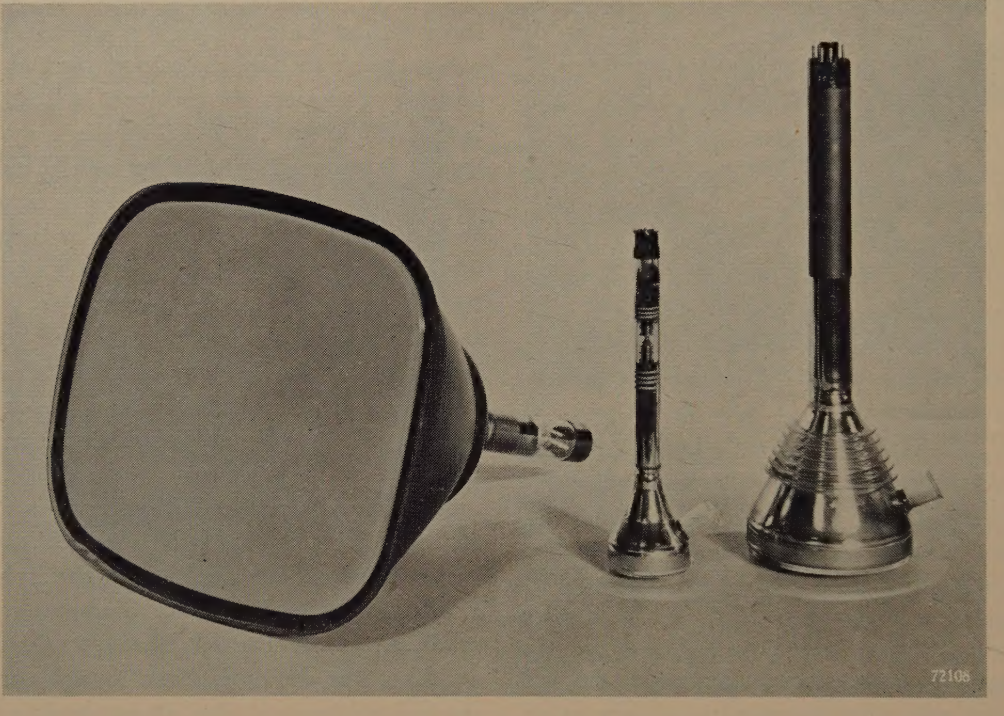


Fig. 12. Cathode-ray tubes for television. At the left a rectangular direct-vision tube with metal cone giving a picture 28 cm × 37 cm. Centre: tube for projection television in the home (max. 1.0 m × 1.2 m approx.). At the right: a tube for projection in halls (3 m × 4 m).

Thus, in order to ensure distortionless conversion of the optical image into electrical signals in these tubes, a number of electronic mechanisms, each of which is already sufficiently complicated, must be made to work together in the correct manner, which means that the manufacture of this kind of tube is certainly no sinecure. As with other types of tube, endeavours are being made to render the camera tube a more compact instrument: amongst other things, this will admit of more convenient

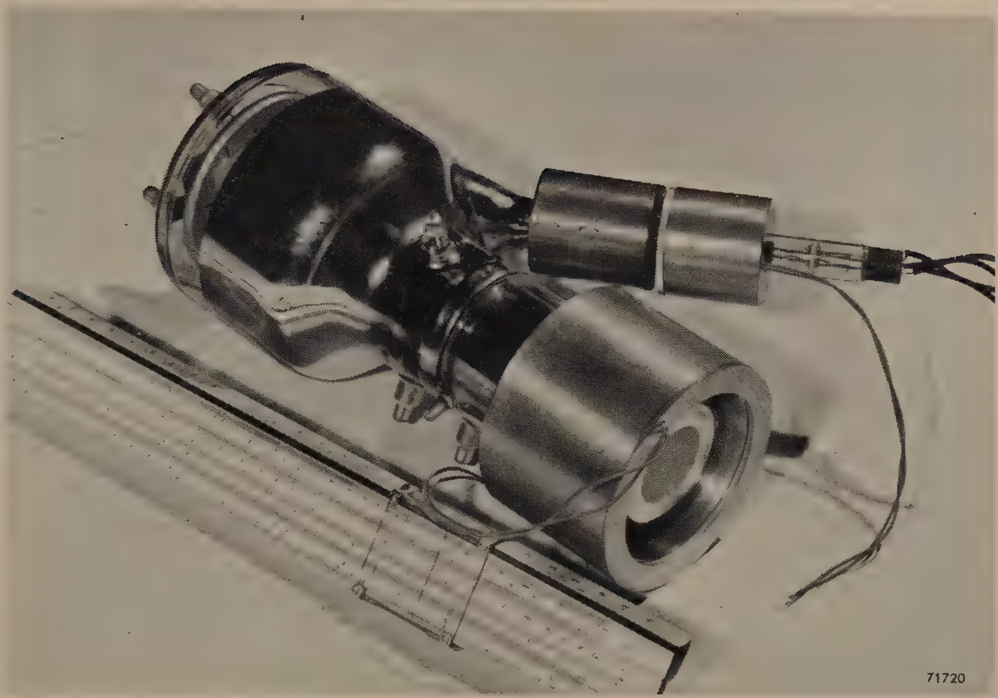


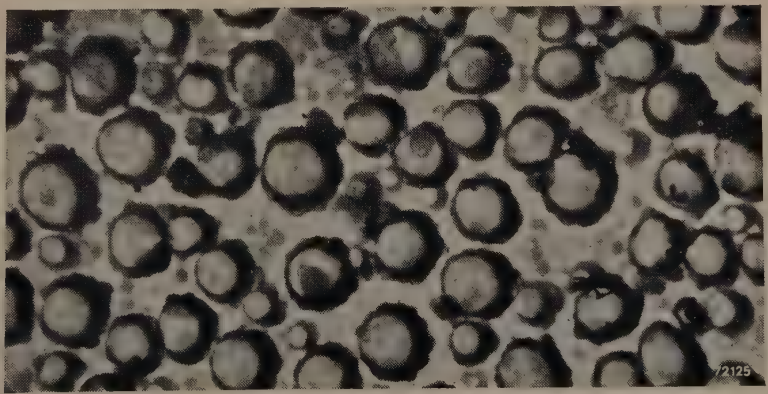
Fig. 13. Image iconoscope. At the near end of the tube will be seen the photo-cathode on which the scene to be televised is reproduced by means of a photographic lens. The large magnet coil shown at this end of the tube produces the field required to focus the image from the photo-cathode on to the mica target, of which a part is just visible at the opposite, wide, end of the tube. At the far side the neck with electron gun and deflection coils for producing the electron beam which periodically scans the target.

proportioning of the optical system, increased depth of focus and less expensive lenses. It is not unlikely that in the future higher sensitivity in this kind of tube will be sought in a wider use of photo-conductivity in place of photo-emission, seeing that the quantum efficiency of the former is in excess of unity. The phenomenon of induced conductivity in insulators also has possibilities to offer in the design of camera tubes [23]).

In conclusion

Only a few examples from the very extensive field of electronic tubes have been discussed here: other important applications, such as those of counter and "memory" tubes, gas-filled tubes, switching tubes, tubes for physical research or for industrial uses and many others have been omitted from our review. In most of these

a



b

Fig. 14. *a*) Photograph of part of the retina of the human eye magnified 750 times. (From S. L. Polyak, The Retina, Univ. Chicago Press, Chicago 1941).
b) Part of the mosaic of an iconoscope, photographed with the electron microscope. Magnification 11 000 ×. The part shown has the same size as that area of the retina enclosed within the small marked rectangle in photograph (*a*).

categories exploratory work is still in full progress and new departures are of almost daily occurrence. The limits of the possibilities offered by electronic tubes are still invisible; even after half a century of research and development, fresh possibilities are continually being discovered.

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LIFE TESTS IN THE ELECTRON TUBE FACTORY

Photograph Walter Nürnberg

THE IMPULSE-GOVERNED OSCILLATOR, A SYSTEM FOR FREQUENCY STABILIZATION

by E. H. HUGENHOLTZ *).

621.396.615.12:621.316.726

The introduction of the quartz crystal in radio technique 25 years ago solved the problem of constructing an oscillator with constant frequency. A drawback of the crystal, however, is that it is suitable for only one frequency so that transmitters which are to operate at different frequencies, require several crystals with the present-day circuit designs, and in a transmitter operating with a continuously variable frequency the crystal cannot be used at all.

This article describes a method by which, with only one crystal, signals can be generated with a series of very stable frequencies, and, if required, a signal can be obtained with a continuously variable frequency which is nevertheless very stable.

Introduction

Two important problems have arisen in relation to the frequency of the oscillators now used in radio transmitters and receivers because of increased traffic and speed. These are:

a) it must be possible to adjust the frequency very accurately;

b) there must be absolute stability of the frequency under changing working conditions, for instance, the ambient temperature or the supply voltages of the valves.

In addition, it is sometimes necessary to change the frequency at short notice.

In transmitters operating at one, constant, frequency and in which frequency changes are comparatively rare, as is the case in broadcasting transmitters, a quartz crystal [1]), usually mounted in a thermostatically controlled container, is generally used. Quartz crystals can also be used in transmitters which use several frequencies for broadcasting, which can be selected as the necessity arises. In that case there must be a crystal for each frequency, and the thermostatically controlled container in which the crystals are mounted must be large. A drawback of this system is that it takes a fairly long time before the required temperature is reached, which may lead to difficulties in mobile installations. The cost of a great number of quartz crystals is another deterrent.

If the frequency of an oscillator must be continuously variable in a certain range, an oscillatory circuit with a variable capacitor is generally used. Various measures are then taken to ensure that the oscillator frequency depends as little as possible upon the operating conditions. Compensation is applied for frequency variations caused by temperature fluctuations, and considerable accuracy of adjustment can be achieved by dividing the required frequency range into a number of smaller ranges which can be selected by changing coils. This often makes the installation fairly complicated, however, and the usual requirements of direct-reading dials cannot be easily met. Moreover, the frequency stability that can be obtained is always less than that which can be achieved with a quartz crystal.

If the frequency range must be continuously variable only in a limited range, a high degree of stability can be achieved with the type of circuit

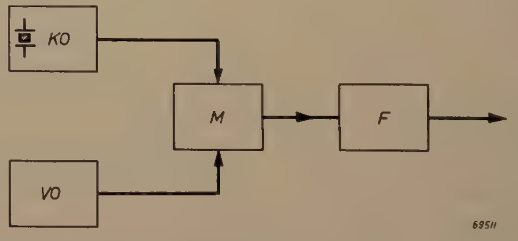

Fig. 1. Fundamental diagram of a system for obtaining an alternating voltage whose frequency can be continuously varied within a limited range and possesses considerable stability.
KO = crystal-controlled oscillator; *VO* = oscillator of which the (continuously variable) frequency is much lower than that of *KO*; *M* = mixer; *F* = filter.

of which the principle is given in *fig. 1*. If, for example, a voltage is required having a frequency that can be varied between 1 and 1.1 Mc/s, it may be obtained by supplying two voltages to a mixer valve *M*. One of these is supplied by a crystal-controlled oscillator *KO*, having a frequency of, for example, 0.8 Mc/s, and the other by an oscillator *VO*, of which the frequency is continuously variable within a range of 0.2 to 0.3 Mc/s. The mixer supplies, amongst others, a voltage at a frequency which is equal to the sum of the two oscillator frequencies

*) N.V. Philips' Telecommunicatie Industrie, Hilversum.
[1]) See, for instance, J. C. B. Missel, Piezo-electric materials, Philips techn. Rev. 11, 145-150, 1949.

and which can be varied between 1 and 1.1 Mc/s. As the frequency changer also supplies voltages with frequencies which are not required, the voltage required will have to be supplied via a filter *F*.

As the oscillator *VO* operates at a comparatively low frequency, a greater frequency stability can be obtained than by using an oscillator having a frequency that is continuously variable between 1 and 1.1 Mc/s. It will be clear, however, that this advantage is no longer present when a voltage is required with a continuously variable frequency over an extensive range, for example, from 1 to 10 Mc/s. This range could be achieved, of course, if a large number of crystal oscillators were available with frequencies of, for example, 0,8 Mc/s, 0.9 Mc/s, 1.0 Mc/s, 1.1 Mc/s etc., and ,in addition, an oscillator with a continuously variable frequency in a limited range, for instance, from 0.2 to 0.3 Mc/s. This solution is obviously not very attractive.

The system described becomes impracticable at higher frequencies when the ratio between the crystal frequency and the frequency of the continuously variable oscillator will generally be large. This means that the relative distance between the required frequency and the unwanted crystal frequency is small, so that high requirements are imposed on the filter *F* which is intended, in the first place, to suppress the crystal frequency (and also the image frequency).

Another possibility is to use one crystal oscillator with a frequency of, for example, 0,1 Mc/s and to obtain from it, by means of frequency multiplication, voltages of which the frequency is a whole multiple of the oscillator frequency. Frequency multiplication may be obtained, for example, by means of a so-called impulse generator, i.e., a generator which does not produce a sinusoidal alternating voltage, but a series of periodic voltage impulses. Such an impulse generator can be synchronized with the crystal oscillator, so that the frequency of the impulses is equal to the frequency of the oscillator voltage. It is possible to obtain from the periodic voltage impulses, by means of filters, alternating voltages of which the frequency is a whole multiple of the impulse frequency. This system, however, requires a complicated set of filters, the very high selectivity requirements of which can hardly be realized, particularly if high ordinals of harmonics are required.

The following will describe another solution of this problem, namely the Impulse-Governed Oscillator (IGO). This method enables the frequency of an oscillator to be adjusted, quite simply, to a large number of values with the same degree of accuracy as would be achieved if a quartz crystal were used for each of these frequencies.

The principle of the IGO is frequency stabilization by means of an automatic control system (often referred to as the "automatic monitor system"), which will be explained first in more detail.

Stabilizing an oscillator frequency by means of an automatic control system ("automatic monitor system")

Fig. 2 shows the principle of this method of frequency control. The stabilized oscillator *SO* supplies a signal of frequency f_1 which is applied to the mixer *M*. A signal of another frequency, f_2, supplied by an oscillator *KO*, and of which the frequency is stabilized by means of a quartz crystal, is simultaneously

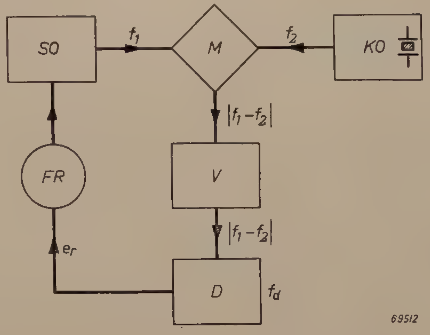

Fig. 2. Fundamental diagram of the "automatic monitor system" for frequency stabilization. *SO* = controlled oscillator, *KO* = crystal-controlled oscillator, *M* = mixer, *V* = amplifier, *D* = discriminator (frequency detector), *FR* = frequency control device.

supplied to the mixer. The mixer *M* supplies, amongst others, an alternating voltage of frequency $|f_1—f_2|$ and this voltage is amplified in the amplifier *V*, after which it is fed to the discriminator (frequency detector) *D*. The circuits of the discriminator are tuned to a certain frequency f_d. The discriminator supplies a direct voltage e_r which is either positive or negative according to whether $|f_1—f_2|$ is greater or smaller than f_d. If $|f_1—f_2|$ is equal to f_d, then $e_r = 0$. The direct voltage e_r, which is called the control voltage, is fed to a control device *FR* which enables the oscillator frequency to be varied. This control device may consist, for example, of a capacitor whose shaft is driven by a small electric motor, or it may consist of a direct control device, for example a reactance valve [2]). The arrangement of *FR* may be such that, when e_r is positive, the frequency of *SO* is reduced and vice versa. Thus, *FR* always influences the frequency of *SO* to such an

[2]) See, for example, Th. J. Weijers, Frequency modulation, Philips techn. Rev. 8, 42-50, 1946.

extent, that $|f_1—f_2|$ is approximately equal to f_d. The frequency f_1 of the oscillator *SO* is, therefore, approximately stabilized to the value $f_2—f_d$ or f_2+f_d, depending on f_2 being larger or smaller than f_1. It is easy to see, however, that this frequency control will never achieve that $|f_1—f_2|$ is exactly the same as f_d, for in that case the control voltage e_r would be zero.

If the discriminator *D* supplies a control voltage e_r of *a* volt per kc/s deviation between $|f_1—f_2|$ and f_d, and if the detuning to which the oscillator *SO* is subjected by the control device *FR* is equal to *b* kc/s per volt control voltage, each deviation of f_1 from the required value $f_2 \pm f_d$ would be reduced by a factor $a \cdot b$ [3]).

The discriminator *D* may be constructed in the same way as the frequency detectors used in F.M.

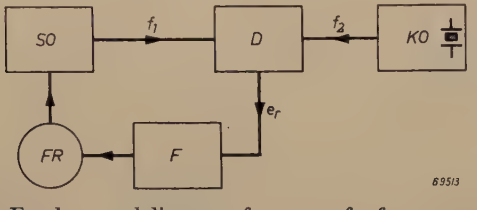

Fig. 3. Fundamental diagram of a system for frequency control. If *D* is a comparative discriminator, a control with proportional action is obtained; if *D* is a phase discriminator, frequency control with integral action is obtained. The significance of the other letters is given in figs 1 and 2.

receivers [2]). The frequency stability of the oscillator *SO* is determined, on the one hand, by that of the crystal oscillator *KO* and, on the other hand, by the stability of the discriminator *D*. Generally speaking, f_d is chosen much lower than f_1 and f_2, so that the absolute value of the residual deviation between $|f_1—f_2|$ and f_d is small in relation to f_1. The error in the frequency of *SO*, caused by the fact that there will always remain a control voltage e_r, however, usually considerably exceeds the sum of the errors of *KO* and *D*.

The same objection is found in another control system of which the principle is shown in *fig. 3*. In this system the voltages supplied by *SO* and *KO* are applied to a so-called comparative discriminator *D*, which is a circuit supplying a control voltage e_r of which the amplitude and sign depend upon the magnitude and the sign of the frequency difference between *SO* and *KO* [4]). Here too, the control voltage is supplied via a low-pass filter *F* to the frequency control device *FR*, which influences the frequency of *SO*. It will be clear that in this case, too, equality of the *SO* and *KO* frequencies cannot be obtained.

It has been explained in the article quoted in footnote [3]) that this disadvantage is inherent in every automatic control system with proportional action. With the aid of a control systeem with integral action, however, the deviation of f_1 with respect to f_2 can be reduced to nothing. To bring about such a control, *D* in fig. 3 must be replaced by a so-called phase discriminator [5]). This refers to a circuit which supplies a control voltage e_r of which the amplitude and sign depend upon the magnitude and sign of the phase difference between the voltages supplied by *SO* and *KO* (thus the frequencies f_1 and f_2 are assumed to be equal in this case). As the frequency is the derivative of the phase [2]), this is a control with integral action for the frequency.

If the frequencies f_1 and f_2 are not equal, *D* will supply an alternating voltage of which the frequency is equal to $|f_1—f_2|$. If this frequency is lower than the cut-off frequency of the low-pass-filter *F*, this alternating voltage is supplied to the frequency control device *FR*, which affects the frequency of *SO*. Provided the inertia of *FR* is not too large, the frequency of *SO* will then vary periodically. When the value f_2 is passed, *D* supplies a direct voltage, which influences the frequency of *SO*, via *F* and *FR*, to such an extent, that this synchronization condition between *SO* and *KO* is maintained. In that case, there is no longer any frequency difference between *SO* and *KO*, but merely a phase shift.

If the frequency of *SO* still differs so much from that of *KO* that the difference is larger than the cut-off frequency of *F*, the frequency stabilization of *SO* will be non-effective. The "collecting zone", i.e., the frequency range in which the frequency of *SO* should be before stabilization begins to act, is, therefore, determined by the cut-off frequency of the filter *F*. The so-called "holding zone" is generally understood to be the frequency range within which *SO* can be detuned (once synchronization has taken place) before synchronism is lost. This frequency range is determined by the range in which *FR* can correct the frequency of *SO* and by the voltage supplied by the phase discriminator *D*, at a certain phase shift between the voltages supplied by *SO* and *KO*. Generally speaking, the "holding zone" is larger than the "collecting zone"

[3]) H. J. Roosdorp, On the regulation of industrial processes, Philips techn. Rev. **12**, 221-217, 1951. This concerns a so-called frequency control with proportional action.

[4]) In order to make a distinction from this, a discriminator, as incorporated in the diagram of fig. 2 and to which only one alternating voltage is applied, is called an absolute discriminator.

[5]) Also discussed in Philips techn. Rev. **13**, 316 (fig. 4), 1952 (No. 11).

Fig. 4 is the fundamental diagram of a system for frequency control with integral action, whereby the frequency f_1 of the controlled oscillator *SO* can be varied over a limited range. This frequency is not stabilized to the frequency f_2 of *KO*, but to a

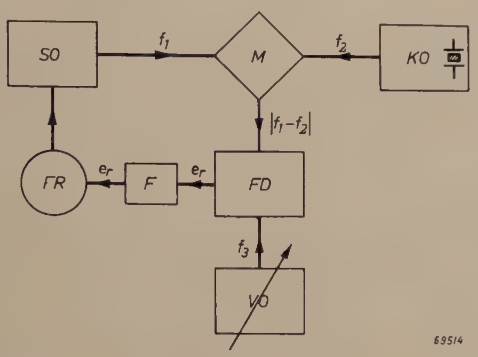

Fig. 4. Fundamental diagram of a frequency control system with integral action, in which there is a difference between the frequencies of the stabilized oscillator *SO* and the crystal-controlled oscillator *KO*, determined by the frequency of the oscillator with variable frequency *VO*. *FD* = phase discriminator. The significance of the other letters is as given in figs 1 and 2.

value which differs by a certain amount from f_2. To achieve this, the voltages supplied by *SO* and *KO* are again supplied to the mixer *M*, which will supply a voltage having a frequency $|f_1-f_2|$. This voltage is applied to a phase discriminator *FD*, together with the voltage from an oscillator *VO* of which

f_1 will be corrected by the same amount. If f_3 is chosen much lower than f_1 (in other words, if f_1 differs but little from f_2), a much larger stability can be obtained, than if *SO* itself were to be constructed as an oscillator with variable frequency.

By making a change in the scheme shown in the diagram in fig. 3, arrangements can be made whereby the oscillator *SO* is synchronized to a harmonic of the voltage of the crystal oscillator *KO*, so that the frequency of *SO* is exactly the same as a whole multiple of the frequency of *KO*.

An impulse generator is used for this purpose, so that such a circuit is known as an Impulse-Governed Oscillator (IGO) [6]).

The Impulse-Governed Oscillator (IGO)

The principle of the IGO is explained with the aid of the diagram of *fig. 5*. The valve B_1 with the oscillatory circuit L_1-C_1, forms an oscillator (Hartley circuit). The frequency of the voltage supplied by this oscillator is influenced by the reactance valve B_2 as well as by L_1 and C_1. An alternating voltage with the oscillator frequency which lags in phase about 90° with respect to the alternating anode voltage, is supplied to the grid of this valve via the phase-shifting network R_2-C_2. This causes the anode current of B_2 to be about 90° lagging in phase with respect to the anode voltage, and the impedance of this valve, between anode and cathode, will

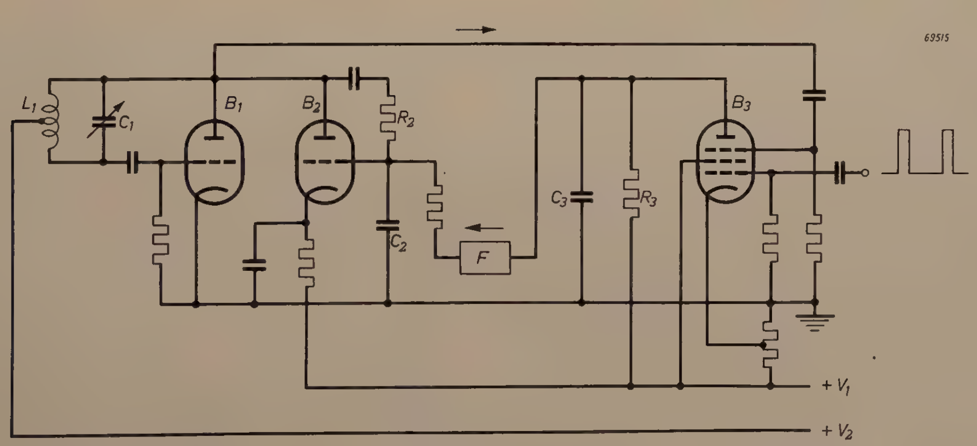

Fig. 5. Circuit diagram of an impulse-governed oscillator. B_1 = oscillator valve, B_2 = reactance valve, B_3 = impulse mixer valve.

the frequency f_3 can be varied within a certain frequency range. The voltage e_r supplied by *FD* is fed via the low-pass filter *F* to the frequency control device *FR*, thus influencing the frequency of the oscillator *SO*. The latter frequency will now be stabilized so that $|f_1-f_2|=f_3$, thus f_1 will be equal to f_2+f_3 or f_2-f_3. If the frequency of *VO* is varied,

have a reactive character. The extent of the equivalent reactance depends, amongst other things, upon the mutual conductance of the valve and hence may be influenced by the grid voltage of B_2.

[6]) See also H. B. R. Boosman and E. H. Hugenholtz, Frequency control in transmitters, Communication News **9**, 21-32, 1947.

The alternating voltage supplied by the oscillator is fed to a so-called impulse mixer valve, the pentode B_3. In this valve, such a large positive voltage is applied to the cathode with respect to the first grid, that no anode current flows in the valve. The first grid is supplied with a series of periodic positive voltage impulses, obtained from an impulse generator which will be described later. The peak voltage of these impulses has such a high value that, during each impulse, anode current flows for a very short time in B_3. The voltage supplied by the oscillator B_1 is applied to the third grid of B_3.

The operation of a circuit in accordance with fig. 5 will be explained with the aid of *fig. 6*. This diagram shows the alternating voltage V_{g3} supplied by the oscillator and the voltage impulses V_{g1} applied to the first grid of B_3. If the frequency of V_{g3} is a whole multiple of that of V_{g1}, the anode current impulses will always occur at the same phasing of V_{g3}.

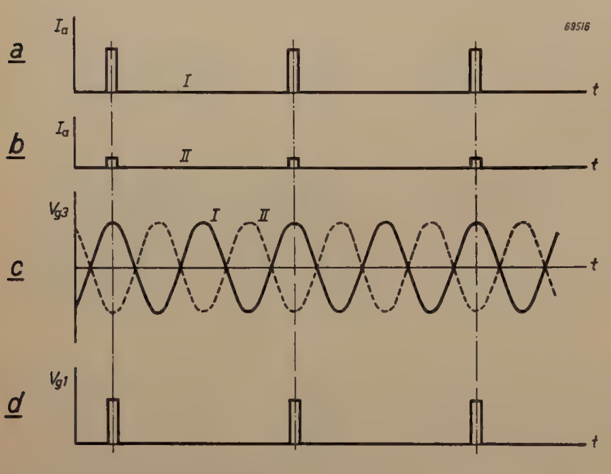

Fig. 6. Graphs explaining the operation of the IGO. The oscillator voltage V_{g3} at the third grid (*c*) and the impulse voltage V_{g1} (*d*) at the first grid of B_3 are presented as functions of time (see fig. 5). The anode current impulses, for the case that the relative phasing of V_{g1} and V_{g3}, is as indicated by *I* and *II* (*a* and *b* resp.), are also indicated as functions of time.

If this is the position indicated in fig. 6*c* by *I*, the amplitude of the anode current impulses will be at its highest, in *II* it is at its lowest. These current impulses cause a periodical charging of the capacitor C_3 (fig. 5). The capacitance of this is so large that a practically ripple-free direct voltage occurs across C_3, and the amplitude of this depends upon the relative phasing of V_{g1} and V_{g3}. This voltage is applied via the low-pass-filter F to the grid of the reactance valve B_2. The average value of this control voltage is obtained when the impulses at the first grid occur at exactly that moment when the alternating voltage V_{g3} passes through zero. Thus, a very small deviation in this setting causes a larger or smaller control voltage to occur. This direct voltage, in turn, influences the frequency of the oscillator, through the reactance valve B_2. If the tuning of this oscillator is changed by an amount which is not too large, the phasing of V_{g3} will change with respect to that of V_{g1}, causing the control voltage to be altered, and the variation in the oscillator frequency is compensated. Thus, the oscillator frequency remains equal to a whole multiple of the frequency of the impulses applied to the first grid.

If this synchronism has not occurred at the moment of switching on, there will be a periodic variation of the voltage produced across C_3, in other words an alternating voltage is obtained [7]) by which, if the frequency is lower than the cut-off frequency of F, the oscillator voltage is frequency modulated. If, during this modulation, the oscillator frequency passes a value which is equal to a whole multiple of the impulse frequency, the oscillator frequency will remain at this value. Theoretically, it should be of no account whether the relative phasing of V_{g1} and V_{g3} adjusts itself in such a way that the impulse lies upon one or the other flank of the sinusoidal curve. According to the direction of the control of B_2, the adjustment upon one of the flanks is, however, unstable and that upon the other is stable.

The low-pass filter F suppresses voltages at the impulse frequency or multiples of it. This prevents the oscillator from being subjected to phase modulation at these frequencies.

When compiling fig. 6, the oscillator frequency was taken to be equal to twice the impulse frequency. It will be clear that synchronization, as described above, may also occur when the ratio between these numbers is a larger integer. The maximum value of this number, as will be obvious from fig. 6, is dependent upon the duration of the impulses. For, if the frequency of the oscillator voltage is so high that the duration of one period is equal to the duration of the impulses, the time integral of the anode current impulses will be independent of the mutual phase position of V_{g1} and V_{g3}, so that, with a phase shift of the oscillator voltage with respect to the voltage impulses, no alteration occurs in the control voltage [8]). For still higher frequencies, a control effect does again occur, reaching a maximum when the ratio of the impulse duration to the duration

[7]) This effect is analogous to the stroboscopic effect in light which occurs, for example, when a rotating wheel with spokes is illuminated by light flashes, whereby the "spoke frequency" deviates a little from a whole multiple of the flash frequency. The wheel is then seen to rotate slowly either clockwise or anti-clockwise.

[8]) A similar problem is described by J. M. L. Janssen, Philips techn. Rev. 12, 52-59, 1950 (No. 2), in particular on pages 57 and 58,

of one period of the oscillator voltage reaches a value of $^3/_2$. When the ratio is 2, there is again no control effect, etc. *Fig. 7* illustrates, on a relative scale, the control voltage as a function of the above-mentioned ratio. Here, the control voltage is the variation of the voltage across the capacitor C_3 at a change in phase of 180° of V_{g3} with respect to V_{g1}. In order to ensure the correct operation of an IGO, the duration of the impulses should preferably

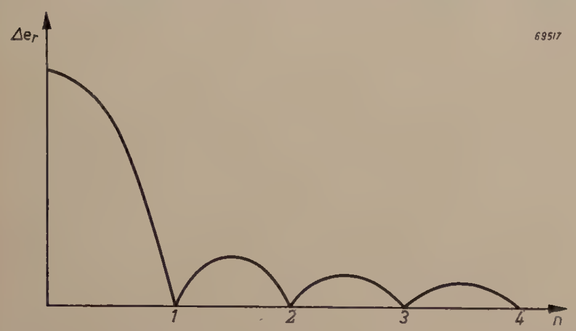

Fig. 7. Variation Δe_r of the control voltage at a phase change of 180° of the oscillator voltage as a function of the ratio n of the duration of the impulse to that of a cycle of the oscillator voltage.

not be larger than half a period of the oscillator voltage. To facilitate the synchronization of an oscillator whose frequency is much higher than the impulse frequency, it will therefore be necessary to make the impulse duration as short as possible. In practice, this is limited by the fact that, at very short impulses, the average value of the current through the valve B_3 (see fig. 5) becomes extremely small. Usually, however, another practical limitation occurs at even lower oscillator frequencies. If, for example, it is required to synchronize the oscillator voltage with the 200th harmonic of the impulse voltage, this means that a deviation of only $^1/_2$% in the proper frequency of the oscillator gives a neighbouring harmonic of the impulse voltage. Hence, high requirements are already imposed on the frequency stabilization of the controlled oscillator itself, in order to maintain the synchronization at the required frequency. For this reason, it appears to be impracticable to make Impulse Governed Oscillators which are synchronized at higher harmonics of the pulse voltage than the 200th.

In an IGO, a distinction can again be made between a collecting and a holding zone, with regard to the frequency range in which the frequency stabilization is operative. The former frequency range is determined mainly by the cut-off frequency of the filter F (fig. 5). For, before the synchronization of the oscillator voltage occurs, the mixer valve B_3 supplies an alternating voltage and only when the frequency of this voltage is so small that it is passed by the filter F, is it possible for frequency stabilization to operate. The "holding" zone is determined by the control voltage, supplied by B_3, when, once synchronization has occurred, the oscillator voltage is subjected to a certain phase shift. Moreover, it depends on the frequency range in which the oscillator can be detuned by the reactance valve. As there is a possibility that the oscillator voltage is synchronized to a wrong harmonic of the impulse voltage, it is desirable not to make the latter frequency range much larger than the frequency interval between two harmonics. By making provision that the holding zone slightly exceeds this interval, it may be achieved that with the (apparently) continuous detuning of the oscillator, the frequency jumps steadily from one harmonic to the next. The construction can then be made in such a way that, by means of a pawl setting, the correct harmonic is selected.

For most applications oscillators must supply a pure sinusoidal voltage with the least possible amplitude and frequency modulation. Deviations from the sine wave may be mainly due to two causes, in an IGO. These are:

a) the filter F (see fig. 5) does not sufficiently suppress voltages with the impulse frequency and, possibly, harmonics of it; this causes frequency modulation of the oscillator;
b) harmonics of the impulse voltage produce an alternating voltage in the oscillator circuit via the mixer valve B_3 (see fig. 5); when a multigrid mixer valve is used, this can occur only via the interelectrode capacitances.

Practice has shown that in the case of frequencies which are not too high, it is comparatively easy to ensure that the voltages with unwanted frequencies are at a level of about —80 dB with respect to the oscillator voltage, mainly because the power of the impulse harmonics is small compared with the output of the oscillator. This level is practically always sufficiently low.

It has been shown that an IGO is a circuit fed with a periodic series of voltage impulses, and that it supplies an alternating voltage of which the frequency is equal to that of one of the harmonics of the voltage impulses. The IGO is, therefore, to a certain extent similar to a filter which passes such a narrow frequency band that only the required harmonic is passed. The frequency to which this fictitious filter has been adjusted may be changed simply by varying the tuning of the oscillator, whereas the width of the frequency band is determined by the

low-pass filter F. For instance, when an interfering alternating voltage is present on the first grid of the valve B_3, in addition to the voltage impulses, and the frequency of this interfering voltage differs very little from that of the required harmonic of the impulse frequency, an alternating voltage will originate in B_3 of which the frequency is equal to the difference of the frequencies of the oscillator voltage and the interfering voltage. It depends entirely upon the filter F, whether this voltage is passed to the grid of the reactance valve or not. If it is, the oscillator voltage will be frequency modulated. If the filter F does not pass the interfering voltage, the oscillator will not be affected by it. It is of importance, in this respect, to choose the cut-off frequency of the filter F as low as possible (unless it is desired to modulate the oscillator frequency). In this case the collecting zone will become small and, in practice, a compromise will have to be accepted.

A circuit like the one depicted in fig. 5 is a form of inverse feedback. If a phase shift occurs in the circuit through which the control voltage of the mixer valve B_3 is supplied to the grid of the reactance valve, there is a possibility that the circuit becomes unstable in this sense that the oscillator does not operate at a constant frequency, but is subjected to periodic frequency changes, thus is frequency modulated. This phenomenon is almost identical to what happens in other circuits with inverse feedback. There is, however, one fundamental difference in the stability condition. An alternating voltage applied to the grid of the reactance valve B_2 gives rise to frequency modulation, whereas the control voltage supplied by the frequency changer B_3 varies through phase shifts of the oscillator voltage. There is a phase shift of 90° [9]) between a phase change (displacement) and frequency change that corresponds with it. Whereas in an "ordinary" amplifier with inverse feedback, a phase shift of 180° is necessary at the critical frequency [10]) to cause instability, here it is merely 90°. In addition, the phase change (displacement) which belongs to a certain frequency change is inversely proportional to the modulation frequency. This means that the higher the critical frequency can be adjusted, the smaller will be the risk of instability in an IGO.

In the example so far reviewed it is assumed that an oscillator is synchronized by an impulse generator. It is, however, also possible to synchronize an impulse generator with an oscillator of which the frequency has been stabilized by a crystal. This may be done by feeding the control voltage supplied by the mixer valve to a control system which influences the frequency of the impulse oscillator. If the frequency of the impulse generator is lower than that of the crystal oscillator, frequency division will take place in which the frequency ratio is the same as the multiplication factor in the previous case.

When an impulse generator is phase modulated, an IGO synchronized to the nth harmonic of it will also be subjected to an n times larger phase modulation, provided the time constant of the control circuit is small enough to pass the highest modulation frequency without much phase shift. Because of the high value of n, which can be reached quite easily in this case, an extensive phase shift can be obtained without much trouble. Thus, a "phase modulation in a narrower sense" [9]) is obtained. In order to get the form of the phase modulation which is usually referred to as frequency modulation, the modulation voltage must be supplied via a network which will correct the frequency characteristic in this sense that the "phase swing" becomes inversely proportional to the frequency of the modulating voltage.

Practical construction of the impluse generator

For most purposes in which an IGO is used, the impulse frequency must comply with high demands of accuracy and stability. The frequency must have a very accurate value, for example, exactly 100 kc/s, so that the IGO is always synchronized to a frequency which is a whole multiple of this value. In view of the fact that, in practice, synchronization to a frequency which is 200 times higher than that of the impulse generator is possible, an IGO may be contructed with a frequency up to 20 Mc/s. As it is preferable for the impulse duration not to be longer than the duration of a half cycle of the voltage produced by the IGO, the maximum impulse time is about 1/40 μsec.

Because of the considerable stability required, an impulse generator is always used, in practice, in which the impulses supplied are synchronized, in their turn, with a sinusoidal voltage. The latter is supplied by an oscillator of which the frequency is stabilized by a quartz crystal, which may be placed in a thermostatically controlled container. *Fig. 8* gives details of a circuit which might be used for this purpose. This is based upon the principle of the ringing oscillator. The hexode section of the triode-hexode B_1 fulfils this function. By a suitable selection of the grid capacitor C_1 and the grid leak R_1, intermittent oscillation is obtained [11]). The setting may be adjusted so that the oscillating is interrupted after a few periods and does not start again until the capacitor C_1, which is charged because of the flowing of a grid current, has been sufficiently discharged via the resistor R_1. Thus, a series of periodic, damped, oscillations occurs in the anode circuit of the hexode section, as is indicated in the lower part of fig. 8. Its repetition frequency is largely determined by the values of C_1 and R_1.

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[9]) See the article mentioned under [2]), p. 44.

[10]) See B. D. H. Tellegen, Inverse feed-back, Philips techn. Rev. 2, 289-294, 1937.

[11]) See J. van Slooten, The functioning of triode oscillators with grid condenser and grid resistance, Philips techn. Rev. 7, 40-45, 1942, and Stability and instability in triode oscillators, Philips techn. Rev. 7, 171-177, 1942.

The triode section of B_1 is used as an oscillator of which the frequency is stabilized by a quartz crystal. The oscillator voltage is applied to the third grid of the hexode section. When the repetition frequency of the damped oscillations is approximately the same as the frequency of the crystal oscillator, synchronization occurs, and these frequencies will become identical. The damped oscillations obtained are now applied to the grid of the valve B_2, which is negatively biased to such an extent that anode current flows only during part of the first half cycle of each of the series of damped oscillations. (For, the amplitude is largest in the first half cycle). Thus, short current impulses occur in the anode circuit of B_2. The coil L_1 has been incorporated in this circuit and with the valve and wiring capacitances it forms an oscillatory circuit. This circuit is periodically excited by the anode current impulses. A second coil, L_2, to which a rectifier has been connected, has been coupled to the first coil L_1, so that this oscillatory circuit can pass on only half an oscillation every time it is excited, thus ensuring voltage impulses of very short duration. These pulses serve to synchronize the actual, adjustable oscillator (see fig. 5).

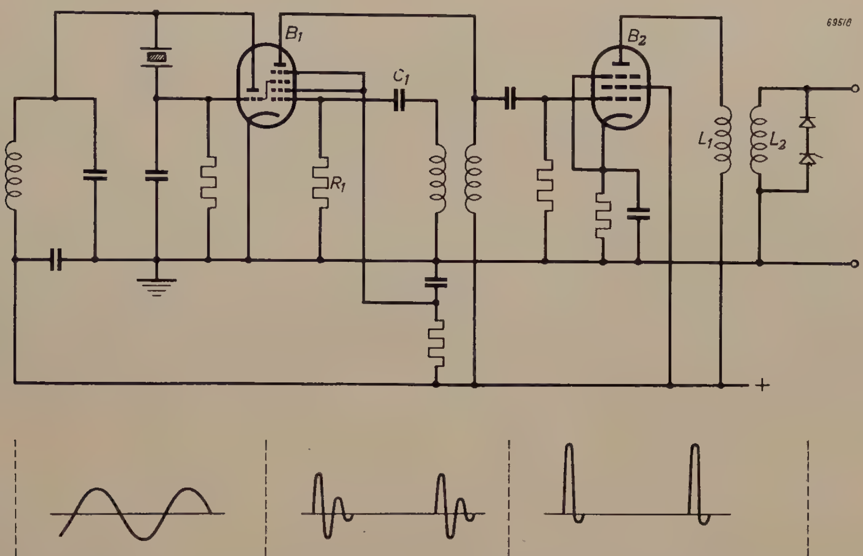

Fig. 8. Diagram of an impulse generator. The triode section of the valve B_1 acts as a crystal-controlled oscillator; the hexode section of this valve forms a ringing oscillator. The impulse duration is shortened by means of the valve B_2. The lower part of the illustration shows the anode currents as functions of the time.

Synchronization of the intermittently oscillating oscillator with the crystal oscillator may also be done so that frequency division takes place. The crystal oscillator may oscillate, for example, at a frequency of 100 kc/s, while the periodic oscillating occurs at a frequency of 50 kc/s.

Fig. 9 illustrates an impulse generator which has been constructed in accordance with the principles described here. The chassis also houses the mixer valve (B_3 in fig. 5). The impulse duration which can be obtained with this device is 1/14 μsec. Proceeding from the condition that the duration of the impulses may, as a maximum, be equal to the duration of a half cycle of the voltage generated by the IGO, it will be clear that this may be used to synchronize an oscillator with a maximum frequency of 7 Mc/s. By adding a third valve to a set-up as shown in fig. 8, connected in the same way as B_2, it has shown possible, in practice, to obtain an impulse duration of 1/180 μsec, so that the synchronization of an oscillator with a frequency of 90 Mc/s is possible.

Decade tuning

It has been explained that an IGO can be synchronized to a frequency which is equal to any multiple of the frequency of an impulse generator, which itself can be synchronized by a crystal oscillator. This makes the IGO admirably suited for use in transmitters, receivers and other apparatus in which so-called decade tuning is used. The principle of this system has already been explained with the aid of fig. 1. An alternating voltage with a frequency, which can be varied over an extensive range, is obtained by supplying two voltages to a mixer valve. One of these voltages is adjustable to frequencies which are a whole multiple of, for example, 100 kc/s. The frequency of the other oscillator is continuously variable in a range of 100 kc/s, for

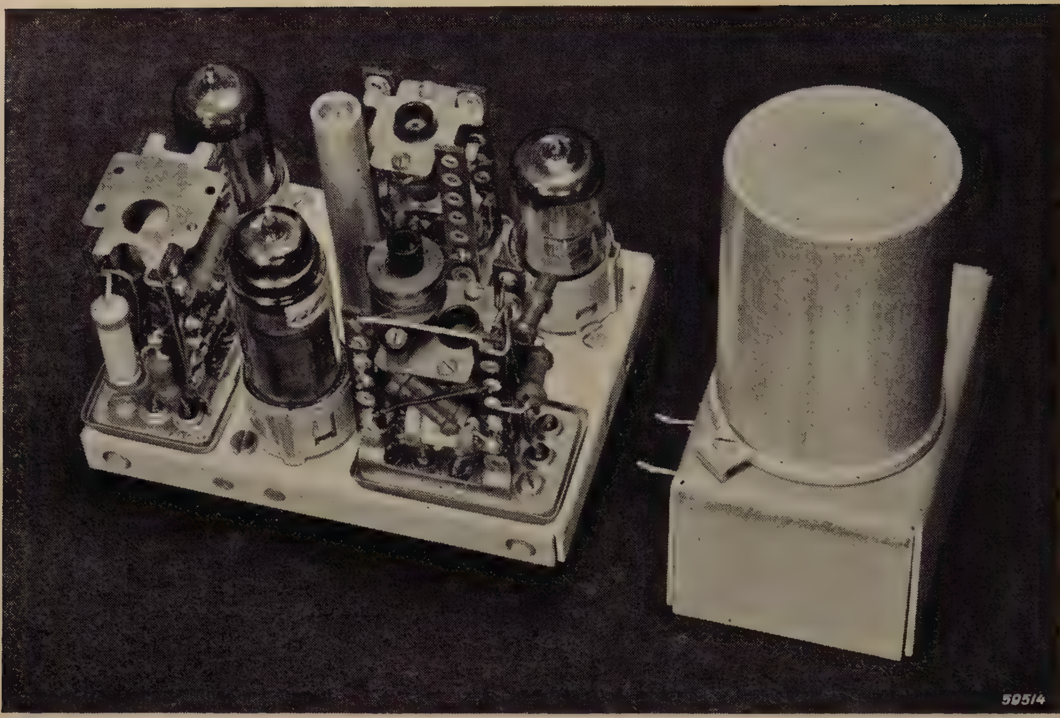

Fig. 9. Crystal-controlled impulse generator and impulse mixer valve (fig. 11: *IG* and *IM*) from the aircraft transmitter SVZ 101.

example, from 200 kc/s to 300 kc/s. The mixer supplies a voltage with a frequency which is equal to the sum of the frequencies of the two oscillators. The stages of 100 kc/s are adjusted by means of a single tuning device, for example, by means of a mechanical pawl system, and a second adjustment device ensures the fine control within the range of 100 kc/s. By placing the two dial controls next to each other, a direct-reading dial is obtained which combines the figures of both scales into one number (*fig. 10*). The accuracy of the readings will be the same for all frequency ranges.

The accuracy with which the required frequency can be adjusted depends upon the frequency accuracy of the two oscillators. The oscillator which is continuously adjustable in the range from 200 to 300 kc/s can be very exact because of the low frequency and the relatively small range. It is much more difficult, however, to construct an oscillator which can be tuned in steps of 100 kc/s and is sufficiently exact at each one of these stages. It will be clear from the above that the IGO is a most attractive means of solving this problem. By synchronizing the impulse generator used with an oscillator of which the frequency has been fixed at 100 kc/s by means of a crystal, the controlled oscillator can be synchronized to every required multiple of this figure and with the same degree of accuracy as that of the crystal-controlled oscillator.

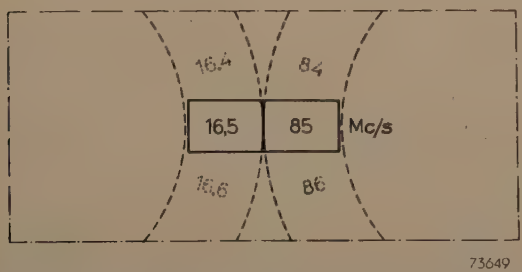

Fig. 10. If the two dials, in decade tuning, are mounted next to each other, a direct-reading scale is obtained on which the figures of both dials are combined (in this case 16 585 Mc/s).

The IGO in an aircraft transmitter

Fig. 11 depicts the fundamental diagram of an aircraft transmitter (Philips Type SVZ 101) which incorporates an IGO. The frequency range of this transmitter is from 2.8 to 24 Mc/s. Twelve frequency bands, each having a width of 100 kc/s, can be selected by means of pawl control knobs [12]).

The impulse generator *IG* has been constructed according to the principle shown in fig. 8. The crystal-controlled oscillator operates at a frequency of 100 kc/s and produces impulses with a frequency of 50 kc/s. The impulse generator and impulse mixer valve *IM* have been constructed as one unit, as illustrated in fig. 9. The IGO which can be adjusted

[12]) A very suitable pawl control knob system for this purpose was described in: W. L. Vervest, Automatic tuning of transmitters, Communication News, **10**, 20-29, 1949.

between 1.55 and 6.1 Mc/s will, therefore, always be synchronized to a frequency which is a whole multiple of 50 kc/s. The alternating voltage supplied by the IGO is used to synchronize the driver stage *ST*. The alternating voltages of the IGO and the driver stage are supplied to the mixer *M* to this end. This supplies, inter alia, an alternating voltage with a frequency equal to the difference between the frequencies of the driver stage and the second harmonic of the IGO. Together with the alternating voltage produced by an oscillator *VO* with continuously variable frequency, the alternating voltage IGO ranges from 1.55 to 6.1 Mc/s, and that of the variable oscillator from 0.2 to 0.3 Mc/s, the driver stage may be adjusted to frequencies between $2 \times 1.55 - 0.3 = 2.8$ Mc/s and $2 \times 6.1 - 0{,}2 = 12$ Mc/s. The output stage *ET* is used as a frequency doubler if higher frequencies are required, so that the highest transmitting frequency amounts to 24 Mc/s.

The variable oscillator *VO* is housed in a small container which is mounted some distance away from the transmitter. *Fig. 12* shows this container. Because of the low frequency and the small fre-

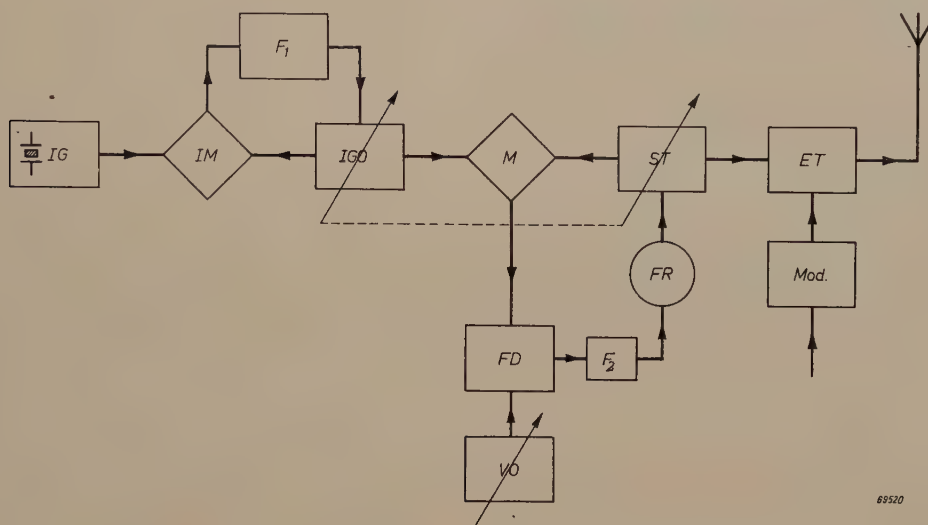

Fig. 11. Fundamental diagram of the Philips aircraft transmitter SVZ 101. *IG* = impulse generator, *IM* = impulse mixer, *IGO* = impulse-governed oscillator, F_1 = low-pass filter, *M* = mixer, *ST* = driver stage, *FD* = phase discriminator, *VO* = oscillator with continuously variable frequency, F_2 = low-pass filter, *FR* = frequency control device, *ET* = output stage, *Mod* = modulation amplifier.

supplied by *M* is applied to the phase discriminator *FD*, which supplies a control voltage which is fed, via the filter F_2, to the frequency control device *FR*, which varies the frequency of the driver stage *ST*. The last-mentioned frequency will always adjust itself to a value which is equal to the difference between the second harmonic of the IGO and the frequency of the variable oscillator *VO*. The last-mentioned is variable between 200 and 300 kc/s, so that the frequency of the driver stage can be varied, by means of *VO*, over a frequency band of 100 kc/s. The driver stage *ST* and the IGO may be adjusted by means of one control knob to the required frequency band, and a padding capacitor maintains the frequency difference between *ST* and the second harmonic of the IGO to about 250 kc/s. Great accuracy is not necessary here, for, as has already been explained, the frequency of the driver stage is exactly determined by that of the IGO and the variable oscillator. As the frequency range of the quency range, the frequency stability of this oscillator is very high. In addition, various other measures have been taken to increase this stability as much as possible. The oscillator valve, for instance, has been coupled to the circuit as loosely as possible and special capacitors have been incorporated to compensate for variations in capacitances due to temperature changes. All elements which are part of the oscillatory circuit, i.e. the coil, the variable capacitor and the trimmers, are contained in a box made of a material which is a bad heat conductor, so that the temperature of these parts will be substantially the same. Because of these precautions, the frequency variation due to a temperature change of 60 °C was limited to 150 c/s.

The frequency control device *FR* influences the frequency of the driver stage *ST* in two ways, namely, both by means of a reactance valve, shunted across the oscillator circuit and by a regulating coil, also shunted across this circuit, and wound

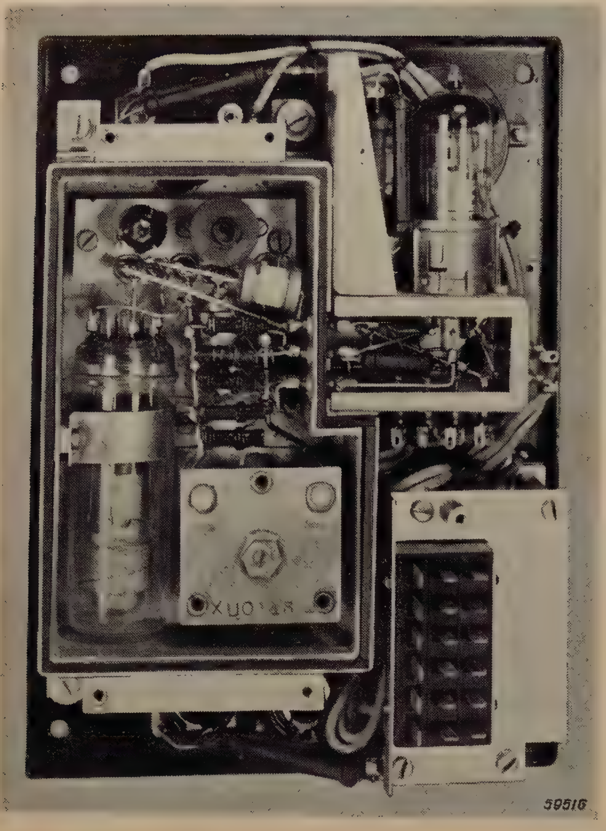

Fig. 12. Control box of the aircraft transmitter SVZ 101. It contains an oscillator with continuously variable frequency between 200 and 300 kc/s.

on a Ferroxcube core. The permeability of this material is varied by the magnetic field of a yoke which is energized by direct current which flows through a coil on the yoke. As the last-mentioned coil has a fairly high inductance, this control is slow, but the frequency variation that can be obtained in this manner is much larger than that which is attained with a reactance valve. On the other hand, control by means of a reactance valve enables compensation of very rapid frequency variations.

Further particulars of the aircraft transmitter SVZ 101 will be found elsewhere [13]).

Other applications of Impulse Governed Oscillators

Finally, there are some other branches of telecommunications in which an IGO might be useful.

In the first place this system offers many possibilities for generating carriers for carrier telephony. In general, these carrier frequencies are multiples of a certain basic frequency, for example 4 kc/s. Each of these frequencies can be generated by an IGO, and these oscillators are all synchronized by means of one impulse generator. This may lead to considerable simplification compared with a system whereby the different carriers are obtained by means of filters from an impulse generator with a frequency of 4 kc/s [14]).

One or more impulse-governed oscillators may, with advantage, be used for generators used for frequency standardization, because it is possible to obtain, from one reference frequency, any frequency with great accuracy. Because of decade tuning, the frequency can be read directly, which practically eleminates errors, interpolations, so often used in other systems, not being necessary here.

By applying the IGO principle in receivers which are constructed according to the so-called double superheterodyne principle, the dial may also be constructed according to the decade system, so that the accuracy of adjustment will be almost the same for all frequencies. The first frequency transformation might be made by an IGO, making frequency steps of 100 kc/s, and the second by means of an oscillator of which the frequency is continuously variable in a range of 100 kc/s.

In transmitters and receivers which have been arranged according to the IGO principle, any frequency up to 30 Mc/s can be found immediately: i.e. there is no "searching". This ensures faster and more reliable traffic. It also makes possible, because of the great choice from very many channels which are close together in frequency (for example a distance between them of only 1 kc/s for telegraphy), the dividing of the traffic over these channels, thus lessening the risk of interference and reducing the inevitable waiting time. This is of utmost importance for aircraft and navigation services.

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[13]) E. H. Hugenholtz, The application of impulse-governed oscillators (IGO) in aircraft transmitters, Communication News **11**, 13-21, 1950 (No. 1).

[14]) See: D. Goedhart and G. Hepp, Carrier supply in an installation for carrier telephony, Philips techn. Rev. **8**, 137-146, 1946.

Summary. This article describes a method of constructing an oscillator which can operate at many different frequencies, yet each possesses the same stability as the frequency produced by a crystal oscillator. Only one crystal is necessary. To do this, the oscillator is synchronized with a harmonic of a periodic impulse (Impulse-Governed Oscillator, IGO), which is obtained from an impulse generator whose frequency is determined by a crystal. The system is similar to a filter which passes a very narrow frequency band and whose tuning frequency can be changed by means of a single circuit.

An IGO permits frequency multiplication and division in a ratio of about 1:200. A combination of an IGO and an oscillator with a continuously variable frequency which is much lower than the frequency of the IGO and which, therefore, possesses a high degree of stability, can be used in a transmitter, which can operate at any required frequency in a certain range and possesses great frequency stability. Other branches of the telecommunication technique, in which an IGO can be used to advantage, are: carrier telephony, frequency standards and radio receivers.

COAXIAL CABLE AS A TRANSMISSION MEDIUM FOR CARRIER TELEPHONY

by H. N. HANSEN *) and H. FEINER *).

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For many years carrier telephone systems have been operated over two different types of cable, namely quadded and coaxial. During this period of rival development, both techniques have gradually improved and both have been successfully adopted by Telephone Administrations in various countries to build extensions of the European long-distance cable network.

This article reviews the history of the development period mentioned above and discusses the comparative advantages of the two systems in a qualitative way. It finally expresses the opinion that neither system has yet achieved a decisive advantage over the other in all fields of application.

Introduction

The justification for carrier telephone systems lies in the economic advantage of providing several speech channels over each physical telephone circuit formed by a pair of line conductors. Before carrier working was introduced, only one audio channel was obtained from each physical circuit, with the result that long-distance telephone systems were composed of large multi-conductor cables whose cost formed a high proportion of that of the entire system. It was, therefore, advantageous to reduce the cable cost by increasing the number of channels per pair of conductors.

At first, in the early 1920's, carrier systems were operated on open-wire lines and, somewhat later, on a few cable circuits. By about 1930 carrier telephone systems with 12 channels were introduced, one such system being operated over each pair of conductors of a paired or (somewhat later) "quadded" cable, the latter containing groups of four conductors arranged in a star. At about the same time, in the United States experiments were being carried out with a type of cable which would admit of several hundred channels per pair, viz. the coaxial cable, containing pairs of concentric conductors. The first experimental multi-channel coaxial carrier telephone system was installed in 1929 at Phoenixville, Penn., by the American Telephone and Telegraph Corp. [1]). About 20 000 miles of coaxial cable are now in operation in the U.S.A.; the transmission system is designed to provide a maximum of 600 channels per coaxial tube. Although the above is only a fraction of the total amount of telephone cable in use, the coaxial cable has clearly proved its worth in a very short time.

Both the above systems were also adopted in Europe. Between 1930 en 1940, coaxial carrier telephone systems were introduced in Great Britain, Germany and France [2]) to work alongside the quad cables, on which 12-channel systems were installed. During and after the second world war, the Netherlands Post, Telephone and Telegraph Service were able to show that on carefully constructed quad cables, the number of channels per pair could be considerably increased, namely from 12 or 24 up to 48 [3]). This discovery had some reactions on the development of the European cable network, but interest in coaxial systems was nevertheless maintained, and steadily increased. In Great Britain a very considerable number of coaxial systems has already been established, and the P.T.T. in France is also engaged on similar activity.

It seems to us to be opportune to present in this Review a brief survey of the rival development of coaxial and quad cables, and to attempt to assess qualitatively their relative advantages and expected fields of application. This will also provide a convenient introduction to future publications describing the work of our laboratories in the coaxial cable field. Two such papers have already appeared [4]).

*) N.V. Philips Telecommunication Industry, Hilversum.

1) Bell Lab. Record **27**, 234, 1949 (Coaxial cable's 20th anniversary). See also L. Espenschied and M. E. Strieby, Systems for wide-band transmission over coaxial lines, Bell. Syst. Techn. J. **13**, 654-679, 1934.

2) Siemens Veröff. a.d. Geb. d. Nachr. **6**, issue of 28th Jan. 1937 (series of articles).
A. H. Mumford, The London-Birmingham coaxial cable system, The Post Office Electrical Engineers' Journal **30**, 206, 270, 1938; **31**, 51, 132, 1938.
R. Sueur, L'évolution de la technique des lignes a grande distance depuis 15 ans, Ann. Télécomm. **6**, 146-164, 1951.

3) G. H. Bast, D. Goedhart and J. F. Schouten, A 48-channel carrier telephone system, Philips techn. Rev. **9**, 161-170, 1947.

4) J. F. Klinkhamer, A through supergroup filter for carrier telephone systems on coaxial cable, Philips techn. Rev. **13**, 223-235, 1952. (No. 8).
H. N. Hansen A new supergroup modulation scheme for coaxial cable telephone systems, Comm. News. **12**, 1-9, 1951 (No. 1).

General description of a carrier telephone system

A carrier telephone system may be classified in three categories, as shown in *fig. 1*, viz.:

the carrier terminal equipment,

the cable,

the line equipment.

The carrier terminal equipment for each physical circuit comprises, at the sending end, a modulation system which enables each channel from the audio-frequency range to be translated upwards in frequency to some specified location in the frequency spectrum. At the receiving end, analogous demodulation equipment translates the incoming high-frequency channels downwards to their original audio-frequency range. The nominal width of each channel is usually 4 kc/s. We shall not deal extensively with the actual modulation process. Usually modulation is done in several steps: By a first transposition to higher frequencies, 12 audio-frequency channels are assembled to form a group; 5 such groups (or less) by a second transposition are combined to form a supergroup. If there are more than 60 channels, each of the supergroups again must be shifted to its final frequency range (the supergroups together form a "hypergroup").

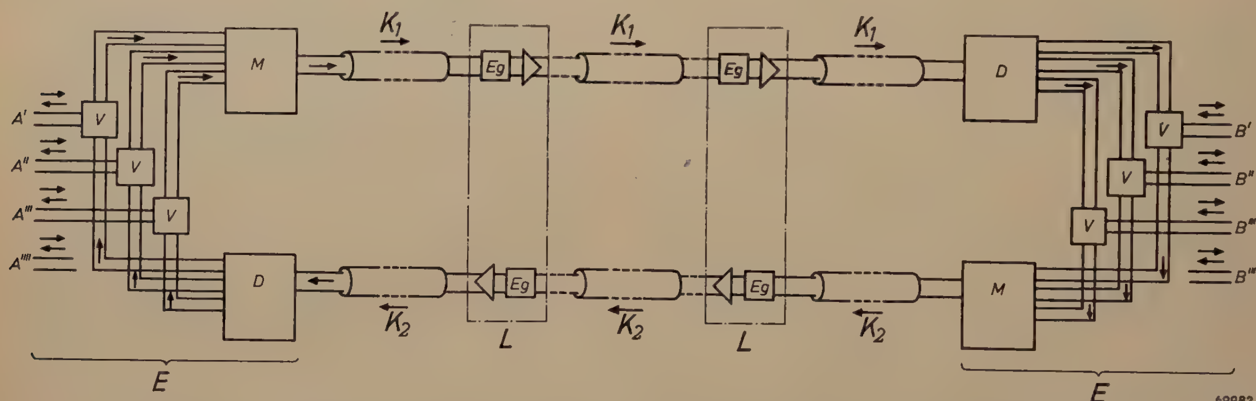

Fig. 1. Block schematic of a carrier telephone system. E = carrier terminal equipment to which the subscriber's lines $A', A'', \ldots B', B'', \ldots$ are connected through 4-wire terminating sets, V. The outgoing signals are passed to the modulating apparatus M and the incoming signals to the almost identical demodulating apparatus D. K_1, K_2 are the "go" and "return" cables. L = line equipment installed in repeater stations. In each station an equaliser Eg and amplifier are provided for each cable pair in each direction.

In quad cable systems, up to the present time it has been normal practice to provide separate "go" and "return" cables for the two directions of transmission.

Line amplifiers must be installed at repeater stations spaced at intervals (repeater sections) along the route; the amplifier gain compensates for the attenuation to which the signals are subjected during their propagation along the cable. In each station, the line amplifier for each physical circuit is associated with an equalising network, so designed that the net overall gain frequency curve of each repeater (by which we mean a line amplifier combined with its equaliser) will match the loss-frequency curve of a repeater section of cable rather accurately. In this way, cumulative linear distortion is avoided, in all channels.

The number of channels per pair

The above-mentioned increase in the number of channels per pair entails a corresponding increase in the frequency bandwidth which must be transmitted over each physical circuit. On cables of the types under discussion, the attenuation increases with frequency. An increase in bandwidth thus leads to an increase in the total amount of gain required throughout the circuit.

This will necessitate a shorter repeater spacing, as can be readily understood. The signal arriving at a repeater must not be attenuated by the preceding cable section to a degree that it will suffer from the thermal noise of the cable and the first amplifying valve of the repeater. Hence, there is a minimum permissible input level. On the other hand, the output level is limited by the power available from the power valves. Thus, the gain to be provided by one repeater is fixed, and a higher total gain can only be obtained by increasing the number of repeaters.

Nevertheless, an increase in the number of channels will economically be justified by the saving in cable cost per channel. Moreover, the number of repeaters *per channel* actually will also be reduced by virtue of the fact that the increase in cable attenuation (i.e. the required gain) is less than proportional to the frequency (it will usually be

found to increase as the square root of the frequency).

There are, however, several limitations which prevent an indefinite increase in bandwidth. Since thermal noise is introduced at the input of every repeater, the minimum permissible level of the signals at the end is dependent on the number of repeaters in circuit. Similarly, the total output power available from a repeater diminishes with increasing frequency; moreover, this power has to be shared between a greater number of channels, so that the maximum attainable level of the signals at the output end is reduced when the bandwidth is increased. The maximum gain obtainable is seriously affected by these effects: for example in a 48-channel system for use on quad cable, the maximum amplifier gain is about 65 dB, but in a 960-channel coaxial system it is not more than 50 dB. Consequently, the repeater spacing must be proportionately shortened to allow for these level limitations, and it is fairly clear that there must be some theoretical optimum beyond which it is not economical to increase the bandwidth.

In systems in use today, operating on either quad or coaxial cables, this economic limit has by no means been reached. The practical limit to the number of channels per pair is imposed by characteristics of the components at present available, in coaxial systems particularly the valves. This limit cannot be raised by any reasonable expedient; for example, it is useless to increase the number of valves per amplifier. These limitations will be explained presently.

Quadded telephone cable

Consider first the common quadded cable, containing pairs of conductors, every two pairs being combined into stars (star quads). *Fig. 2* shows a 12-quad cable of the type widely employed by the Netherlands P.T.T. Air-spaced paper insulation is adopted, and the cable is covered by a lead sheath as a protection and screen, and to prevent the ingress of moisture. Further protective layers are provided to prevent physical damage to the cable if it is laid directly in the ground.

The cable attenuation is shown as a function of frequency in *fig. 3*. The absolute value of the attenuation, other things being equal of course, depends on the diameter of the conductors. The choice of conductor diameter, within limits imposed by electrical and mechanical requirements, is governed by economic considerations again; large diameters result in costly cables, but reduced attenuation and, therefore, fewer repeaters. The most economic choice will be greatly affected by the market prices of copper and lead. In practice it is found that the optimum conductor diameter lies between 0.9 anp 1.3 mm. The attenuation at 200 kc/s is about 2,4 dB/km for 1.3 mm conductors. Assuming a repeater gain of 60 dB, the repeater spacing for such a cable is 25 km, which corresponds with normal practice applicable to the 48-channel system used

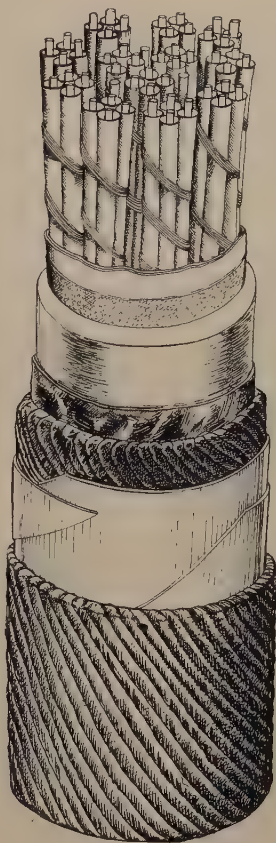

Fig. 2. Typical 12-quad cable. Each star quad contains two diagonal pairs. Air-spaced paper insulation is used. The cable is sheathed with lead. Outer servings and steel tape armouring are applied over the sheath.

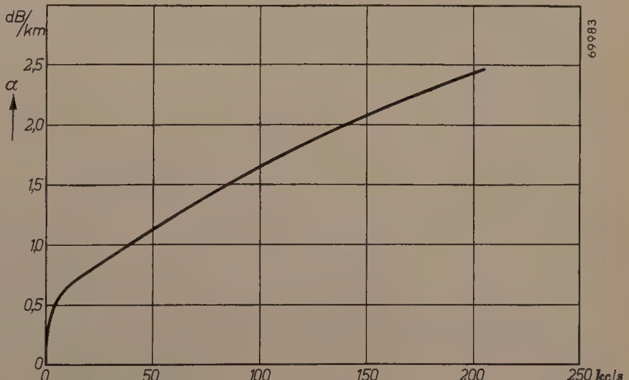

Fig. 3. Attenuation-frequency characteristic of a 1.3 mm pair in a quad cable. The diagram shows the attenuation α in dB/km as a function of frequency f (in kc/s).

in the Netherlands and elsewere (frequency band 12-204 kc/s, repeater spacing about 26 km).

So far, it has not been possible to increase considerably the number of channels per pair in quad cable due to limitations of crosstalk. Crosstalk — the transfer of signals between different pairs of conductors — is due to the effects of capacitance unbalance and mutual impedance between the various pairs in the cable. Crosstalk in quad cables increases with the frequency of the signals. By suitable precautions in the manufacture and installation of modern quad cables [5]), it has been found possible to keep the crosstalk within permissible limits up to a frequency of 250-300 kc/s, which is sufficient to allow the operation of 60-72 channels per pair. Possibly some further progress may yet be made, but the most optimistic estimate of what may be achieved in the future does not exceed 120 channels per pair.

If such an increase in the number of channels per pair is ever to be attained, a very efficient balancing technique will have to be evolved. For this purpose, the introduction of variable balancing elements at intermediate points within each repeater section is contemplated.

Attention today is concentrated on possible ways of diminishing the crosstalk because new methods of cable construction promise an increase of frequency bandwidth without a corresponding reduction in the repeater spacing. The attenuation of present types of cable is partly attributable to dielectric losses in the paper insulation, especially at frequencies in the order of 200 kc/s, the loss angle tan δ then reaching a value of 120×10^{-4}. Various manufacturers are developing new types of cable in which the insulating material consists of polyethylene, styroflex or polystyrene foam, whose low dielectric constant and loss angle (tan $\delta = 3 \times 10^{-4}$ or less) enable the HF attenuation to be reduced considerably below the figures applicable to paper cable. In fact, the attenuation of foam-insulated cables at 550 kc/s is comparable with that of paper-insulated cables at 250 kc/s.

Coaxial telephone cable

A coaxial pair or tube generally consists of a copper wire which forms the inner conductor or core, supported by some insulating structure, so that its axis coincides with that of a tubular outer conductor. The outer conductor is usually also of copper, but is sometimes made of aluminium, which is cheaper. Several such pairs or tubes may be laid up together with auxiliary conductors to form a coaxial cable. The auxiliary conductors are required to provide a service telephone, extended alarm circuits etc. *Fig. 4* shows an 8-tube coaxial cable; a type has been standardised in the United States. The diameters D and d of the outer and inner conductor respectively in these cables, both being made of copper, (*fig. 5*), bear roughly the ratio:

$$\frac{D}{d} = 3.6 \qquad (1)$$

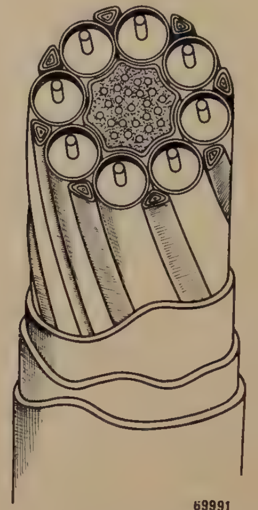

Fig. 4. Typical coaxial cable as used in the USA. Eight coaxial tubes are laid up helically over a core consisting of auxiliary signal wires. Similar sheathing and armouring as fig. 2.

In one cable standardised internationally, d = 2.6 mm and D = 9.4 mm (3/8″). The inner conductor is supported along the axis of the tube by discs or beads of polythene or other materials, spaced 1″ apart, or alternatively by helical band of polythene or styroflex, etc. The remaining space is filled with dry air or gas, sometimes under pressure.

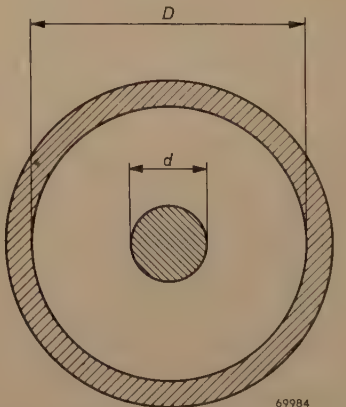

Fig. 5. Cross-section of a coaxial tube. The ratio of diameters D/d is usually 3.6.

The motivation of the condition (1), as well as a brief summary of the transmission characteristics of coaxial cables, will be found in the Appendix to this article.

We can now state in a few words the reason why it has been possible to employ such a very much greater number of channels per pair in the coaxial cable than in the quad cable: *in the coaxial cable crosstalk*

[5]) See L. J. E. Kolk, Het balanceren van draaggolfkabels voor 48-kanalen systemen, PTT Bedrijf 3, 59-74, 1950 (in Dutch).

between pairs does not increase, but decreases with frequency.

This advantage can be accounted for by a consideration of skin effect. At sufficiently high frequencies, the signal currents flow nearly entirely on the outer surface of the inner conductor and on the inner surface of the outer conductor. The outer surface of the outer conductor carries practically no current, and between two points on this surface only very small alternating voltages will occur. When considering the influence of these voltages on an adjacent tube, the skin effect again is of importance: the density of the currents induced in that tube is greatest at the outer surface of the outer conductor, it sharply decreases towards the interior, which means that the interference is hardly capable of penetrating to the transmission circuit proper. This "shielding action" of the outer conductor of each tube is most effective at high frequencies. Thus, at sufficiently high frequencies, crosstalk will become entirely negligible.

The crosstalk shielding effect of the outer conductor not only permits the use of very high signal frequencies, but it also makes it possible to place the coaxial tubes, used for opposite directions of transmission, within a common lead sheath. Owing to the large difference in level between incoming and outgoing pairs at repeater stations, so exacting requirements as to crosstalk must be met with quad cables so that in this case separate "go" and "return" quad cables must, in general, be used.

From the above it is apparent that crosstalk on coaxial cables is only serious at the lowest frequencies transmitted. The lower frequency limit, therefore, is chosen fairly high. If it were desired to extend downwards the lower limit of the frequency band transmitted, this could be done by increasing the thickness of the outer conductor, but only at the expense of increased cable cost. Some quantitative idea of this relationship may be obtained by a consideration of the depth of penetration ϑ of the signal and crosstalk currents into the outer conductor, i.e. the depth within the conductor at which the current density has diminished by a factor $1/e$. For a conductor having a resistivity ϱ (in ohm metres) and a relative permeability μ_r, we may write:

$$\vartheta = \frac{10^5}{2\pi}\sqrt{\frac{\varrho}{\mu_r f}}\ \text{mm} = \frac{C}{\sqrt{f}}\ \text{mm}. \quad (2)$$

In the above expression, the frequency must be expressed in kc/s. For copper it is found that $C \approx 2.1$; for aluminium $C \approx 2.75$. In order to reduce the depth of penetration to a few tenths of a millimeter, i.e. to an amount small compared to the minimum thickness of tape which can be applied without excessive mechanical difficulties, the frequency must exceed 100 kc/s. In coaxial systems on multitube cables, the lowest frequency transmitted is never less than 60 kc/s, often more [6]).

The upper limit of frequency in coaxial systems so far installed is approximately 2.8 Mc/s (660 channels).

Standardisation by the C.C.I.F. will in due course extend this band up to 4 Mc/s (see *fig. 6*), and work is now in progress on a transmission band up to 8 Mc/s. Naturally, at these increased frequencies the attenuation per km of cable is higher, and it will be necessary to use shorter repeater sections than have hitherto been customary. Nevertheless, as indicated above, this is not necessarily an economic dis-

[6]) From equation (2) it will be seen that a high permeability results in a smaller penetration depth. For iron, $C \approx 0.3$. To save copper, the outer conductor therefore is sometimes covered with steel tape.

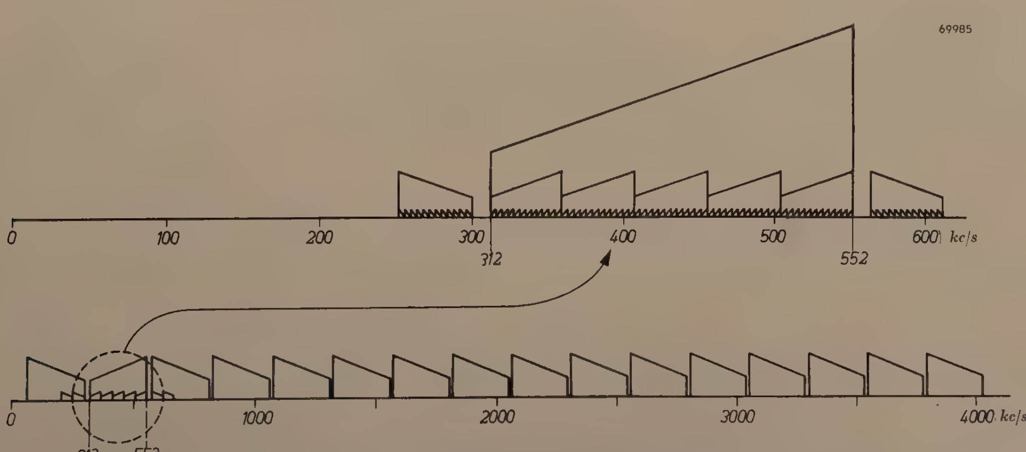

Fig. 6. Frequency allocation of a 960-channel coaxial cable system with 16 supergroups. The part within the dotted circle (supergroup no. 2) is enlarged to show more clearly the 5 groups per supergroup and the 12 channels (of nominal bandwidth 4 kc/s) in each group.

advantage, since the attenuation only increases as the square root of the frequency. For present types of coaxial cable the attenuation is given by the following formula:

$$a = \frac{0{,}07}{D} \sqrt{f} \text{ dB/km}, \quad . \quad . \quad . \quad . \quad (3)$$

where f is again expressed in kc/s and D is the inside diameter of the outer conductor in cm. *Fig. 7* shows the attenuation-frequency characteristic of a cable in which D has the standardised value 0,94 cm, mentioned above. The attenuation is 4.8 dB/km at 4 Mc/s. Assuming a repeater gain of 50 dB (which allows a suitable margin), the necessary repeater spacing is thus about 9.5 km.

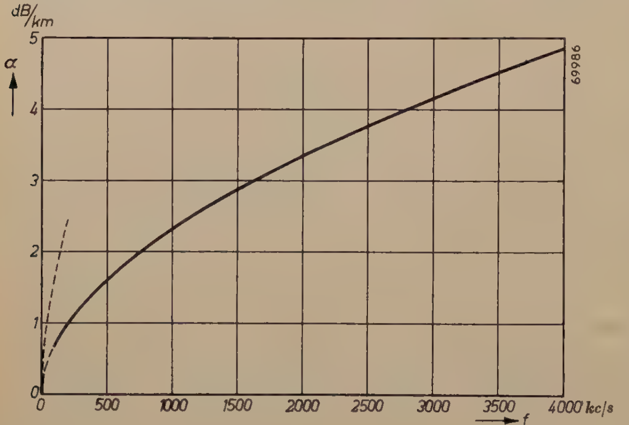

Fig. 7. Attenuation-frequency characteristic of a coaxial tube; D, the inner diameter of the outer conductor, is 0,94 cm. The broken line shows the corresponding attenuation of a 1.3 mm pair in a quad cable, reproduced from fig. 3.

There is no economic objection to a considerably higher maximum frequency and a still smaller repeater spacing, but further progress in this direction is inhibited by technical difficulties in the repeaters.

Characteristics of repeaters for coaxial systems

Fig. 8 shows a comparison between two repeatered cable systems of equal length, of the 12-quad (i.e. 24 pairs) and coaxial types respectively. "Go" and "return" paths are shown in each case, In the case of the quad cable it is assumed that each pair transmits 60 channels, i.e. 1440 in all, and that the repeater spacing is about 23 km. The coaxial tube accommodates 960 channels, the repeater spacing being taken as 9.5 km.

It is interesting to observe the saving in line amplifiers and physical circuit material per channel shown in the second diagram. Taking into account "go" and "return" transmission paths, the quad cable requires $2/(60 \times 23) \approx 1/700$ amplifier per channel km, whereas in the second case the corresponding figure is $2/(960 \times 9.5) \approx 1/(4500)$ amplifiers per channel km.

It is not our aim, however, once again to insist on the economic advantage of one system over the other. (To do this objectively we should have to consider the prices of cable and repeaters as well as those of the terminating equipment). We wish

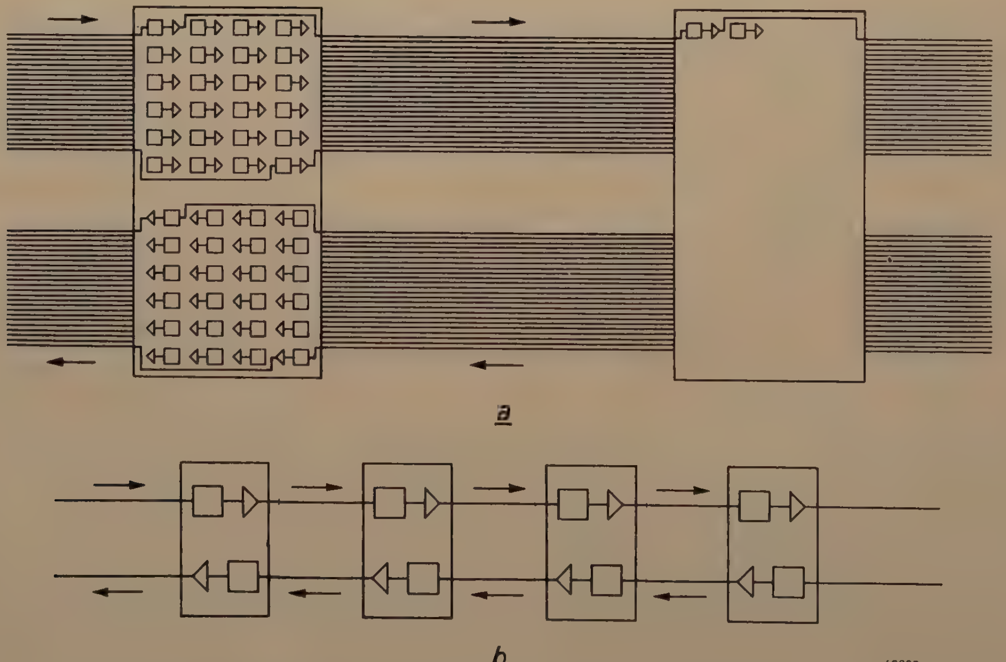

Fig. 8. *a*) Block schematic of go and return 12-quad cable with repeaters. Total number of channels is $60 \times 24 = 1440$.
b) Block schematic of a pair of coaxial tubes transmitting 960 channels.

rather to draw attention to some other features of this diagram:

a) In the coaxial system each repeater handles 960 channels, whereas in the quad cable case, the corresponding figure is 60 channels.

b) In the coaxial system each bi-directional physical circuit is equipped with 6 repeaters in a distance of 23 km, whereas in the quad cable system there are only 2.

c) Finally, in the coaxial cable there are many small stations, and in the quad case, fewer large ones.

In consequence of these differences the line apparatus for coaxial cables in several respects must satisfy more stringent requirements than that used for quad cable. First, the reliability should be considered. The breakdown of one coaxial repeater would disable 960 channels simultaneously, which in telephony work would be regarded as little short of a catastrophe. Each repeater must therefore be provided with the necessary spare equipment, and it might even be advisible to provide a pair of coaxial tubes, fully equipped, as a standby.

Secondly, the stability is of importance. In a long-distance connection each channel passes through numerous repeaters in tandem. To maintain a constant level of all channels at the receiving end, variations of the individual gain of each repeater are strictly limited. Since there are more repeaters in tandem in a coaxial than in a quadded cable, for circuits of equal length, the limits for a coaxial system must be more stringent. This necessity is enhanced by the fact that the coaxial system is generally used for longer distances than quad cables.

Thirdly, non-linearity in the repeaters, especially the valves, gives rise to intermodulation, i.e. a form of (un-intelligible) crosstalk between channels. Since many channels contribute to intermodulation into any given channel, this effect is apparent as a background noise which is dependent upon the total number of channels per pair and upon the number of amplifiers in tandem. In both these respects coaxial cable again is at a disadvantage compared with quad cable.

The linearity of the amplifiers and their constancy of gain can theoretically both be improved by the use of negative feedback. In our case this requires a sacrifice of gain of 30-40 dB in each repeater. This, however, is particularly inconvenient because the stage gain of a broadband amplifier capable of handling 960 channels is already relatively small. This technical difficulty even sets a fundamental limit to the number of channels which can be transmitted over a single tube. In fact, the parasitic capacitances of amplifier valves, input and output transformers, etc., which are responsible for the decrease of stage gain at increasing bandwidth, also give rise to a phase displacement which increases with frequency; for this reason, there is a limit to the number of stages across which negative feedback may be applied; the limit in this case is three stages. It might be suggested that an amplifier could be built, for example, with two units, each consisting of three stages, with negative feedback applied to each unit individually. The sacrifice of 30-40 dB gain, however, would then apply to each such three-stage unit. Obviously, the placing of units in tandem will be of no use and no gain at all can be obtained if the available overall amplification per three-stage unit is already less than 30-40 dB. Beyond this limit any increase in the bandwidth transmitted over coaxial systems is thus conditional upon improvements in the characteristic of the components, particularly in respect of the capacitance and slope of the amplifier valves [7]).

These, then, are the principal factors which for the moment set a limit to the increase of the frequency bands employed on coaxial cables.

The dispersal of the line apparatus over a large number of small stations, as demonstrated in fig. 8, is, unlike the subjects previously discussed, not primarily a circumstance which limits bandwidth; but an increase of bandwidth would certainly add to the practical difficulties of feeding power to a large number of repeater stations, possibly containing an increased number of valves. In quad cable systems it is customary to feed the line equipment from a local mains supply, a standby power supply being provided by a battery or engine-generator set. It is, however, not economical to furnish each of a large number of closely spaced coaxial repeater stations with its own local mains supply, nor with an emergency power source. The best solution of this problem so far devised is to feed each repeater station from a power supply transmitted over the cable itself. The necessary A.C. power is fed into the cable at stations spaced at intervals of up to 300 km along the cable route; these stations are provided with alternative sources of power supply, as a precaution against mains failure. By centralising the standby plant in this way, the problem of sup-

[7]) For practical feedback and amplification values for a given bandwidth, see H. W. Bose, Network analysis and feedback amplifier design, Van Nostrand, New York, 1945. See also J. te Winkel, A note on the maximum feedback obtainable in an amplifier of the cathode-feedback type, Philips Res. Rep. **5**, 1-5, 1950. In this connection it may be said that the use of Ferroxcube for the cores of input and output transformers in the amplifiers already represents an important step forward.

plying power to a large number of small stations has been solved in a very simple way and at a low cost per repeater. This system is all the more attractive since the solid copper core of the coaxial tube, which, with the exception of the extreme outer layer, does not serve the transmission of the signals, is then effectively utilized. Another advantage of the method is the easy solution it offers for feeding repeater stations in sparsely inhabited country where there may be no electric mains for hundreds of miles.

Another important problem in the line equipment, especially of coaxial systems, is the variation of a, the attenuation of the cable, as a function of temperature T. Evidently, the factor preceding $\sqrt{f}$ in equation (3) and in which among other things the resistivity of the copper is involved, will vary with temperature. Now, $\mathrm{d}a/\mathrm{d}T$ will be obtained by multiplying the temperature coefficient of that factor — whatever function that may be — by $\sqrt{f}$. This means that fluctuations in the cable temperature will affect the level of the high-frequency channels considerably more than the low-frequency ones. The variations in level in coaxial systems have been reduced to reasonable proportions by providing an efficient form of automatic level regulation controlled by the line pilot signals. The extent of such variations for the reason explained being considerably less on quad cable systems, less elaborate regulation in that case is satisfactory.

The problem of interconnection

To complete our consideration of the relative merits of carrier systems on coaxial and quad cables, we shall now examine their respective adaptability to meet the demands of practical handling of carrier systems, in other words their "flexibility". Carrier telephone systems usually form part of a national or international telecommunication system. At various intermediate towns or junction points along the route a sufficient number of channels must be terminated to accommodate local traffic requirements. This number will vary over a wide range, and may not be constant. Temporary changes in the distribution of channels may be necessary on account of important events; or perhaps there may be steady growth of traffic due to an increase in the number of subscribers or the introduction of automatic operating procedure on trunk circuits. The structure of a trunk system is thus always changing and demands a certain degree of adaptability of the carrier system.

Systems incorporating 12-60 channels per pair such as those operated over quad cables, fulfil this condition admirably; by terminating a suitable number of pairs at each intermediate town or junction point, the distribution of channels can be readily adjusted to meet traffic requirements. On the other hand, the same technique cannot be applied so easily to the coaxial system. Terminating one pair of coaxial tubes (go and return) at an intermediate town would mean providing for this town a hypergroup containing many hundreds of channels (potentially 960), i.e. much too large a unit for convenient use. This is the price which has to be paid to secure the economic advantages of transmitting a very large number of channels on a single physical circuit.

Some months ago this was reviewed in detail in this journal [4]) and reference was made to the possibility of interconnecting two coaxial systems on a "through group" or "through-supergroup" basis, so that one hypergroup is no longer to be regarded as an indivisible unit, but some channels can be terminated at an intermediate point and the others can pass through in transit. The through supergroup procedure requires that the incoming hypergroup be demodulated to its component basic supergroups, whose frequency range is 312-552 kc/s. The supergroups which terminate are demodulated completely to audio frequency, whereas those in transit pass via "through-supergroup filters" to the supergroup modulators, to take their place in the outgoing hypergroup.

Although in principle this arrangement imparts to the coaxial system the flexibility of 12- or 48-channel quad cable systems, the additional equipment at the transit point inevitably must detract from the economic advantage of the coaxial system. An alternative and simpler method of providing transit facilities exists; this is known as echelon working and it consists in allocating to the traffic between each pair of towns in a coaxial cable network the exclusive use of one or more of the supergroups provided by a pair of tubes. The intermediate carrier terminals then only need be connected in parallel with the main transmission paths by any convenient method, without any of the special filters mentioned above. Admittedly the available bandwidth in general is not fully utilised when this scheme is adopted, and consequently part of the essential feature of the coaxial cable system is sacrificed.

A completely different solution of the problem that is being studied involves a rearrangement of the supergroups with a wider mutual separation. This proposal will usually result in a less efficient utilisation of the frequency bandwidth, but will permit of a new technique, the use of "band-stop filters" to suppress one or more of a range of supergroups for purposes of extraction at intermediate terminals. This would greatly simplify the distribution problem [4]).

The above discussion indicates that coaxial and quad cables each have their own characteristics which adapt them for specific applications, neither type having a decisive advantage over the other in all fields. Future developments may eventually lead to one system gaining ground at the expense of the other. On the basis of present technique, the coaxial system undoubtedly is to be preferred for projects in which a large number of channels must be provided over long distance, and in which the characteristic economy in the cost of cable and repeaters is consequently obtained. In sparsely populated areas where power supplies are mostly non-existent, the coaxial system has the important advantage of power feed over the cable; in some areas, e.g. in the U.S.A., this advantage has been decisive. For comparatively short distances in thickly populated country, particularly where the cable network forms a closely triangulated pattern, quad cable systems with relatively few channels (e.g. 48) have generally been preferred on account of their greater flexibility. There are many marginal cases between these two extremes; the ultimate choice will here depend on the progress of future development, for example the increase in frequency bandwidth of quad cable mentioned above, and the proposal to feed power over quad cable to repeater stations. These expected improvements will then have to be compared with the corresponding evolution of coaxial technique.

Finally, it should be mentioned that the feasibility of relaying television programmes is being advanced as an important argument in favour of coaxial cable. For this purpose a bandwidth of at least 4 Mc/s is required. A combination of telephone and television transmission over the same medium would undoubtedly be attractive.

Appendix: Transmission characteristics of coaxial cable

The transmission characteristics of any line, for a signal of frequency $f = \omega/2\pi$, are determined by its resistance R, inductance L, leakance G, and capacitance C, which properties are continuously distributed along the line and which are expressed per unit length. The values of R, L, G and C, the "primary constants" are usually functions of frequency, in spite of their name, but only to a slight extent in the case of L and C.

The theory of transmission becomes fairly simple when the line is uniform and when $\omega L \gg R$ and $\omega C \gg G$, which is always the case in practice at high frequency, i.e. well beyond the audio range. According to theory, in that case the ratio of signal voltage E to signal current I, both assumed sinusoidal and transmitted into an infinitely long line, is the same in all points and equal to $\sqrt{L/C}$; this ratio is independent of frequency and is called the characteristic impedance. The losses in transmission are caused by the resistance R and the leakance G. If the current and voltage transmitted into an infinite line are I_1 and E_1, at unit length from the sending end the smaller values I_2 and E_2 are found. The attenuation of the cable is defined by the following equations:

$$\alpha = \log_e |E_1/E_2| = \log_e |I_1/I_2|.$$

From theory, the following simple formula for α is obtained:

$$\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} \quad \dots\dots\dots \quad (4)$$

It contains two terms, corresponding with the two distinct causes of loss: the first term represents resistive attenuation due to R, and the second term, always considerably smaller, the leakance attenuation due to G.

So far the theory applies to all types of line. In order to apply formula (4) to our case, we need to relate the primary constants to the dimensions and material constants of the coaxial cable.

Consider first the resistance R. The high-frequency current chiefly flows in a thin cylindrical layer at the outer periphery of the inner conductor (diameter d) and at the inner periphery of the outer conductor (diameter D). Now R can be taken as equal to the D.C. resistance of two tubular conductors with the diameters d and D respectively and a wall thickness equal to the penetration depth ϑ. The cross-sectional areas of these conductors are $\vartheta \cdot \pi d$ and $\vartheta \cdot \pi D$, and according to eq. (2) we have $\sqrt{\varrho_1/f}$ and $\sqrt{\varrho_2/f}$ respectively, ϱ_1 and ϱ_2 being the resistivities of the inner and outer conductor material respectively. The resistance R then is proportional to

$$\sqrt{f}\left(\frac{\sqrt{\varrho_1}}{D} + \frac{\sqrt{\varrho_2}}{d}\right).$$

The capacitance C as well as the inductance L of two coaxial cylinders depends on the ratio of the diameters only, not on their absolute values. It is found that $\sqrt{L/C}$, the characteristic impedance, is proportional to $\log D/d$. The resistance part of the cable attenuation given in equation (4) then becomes proportional to

$$\frac{\sqrt{f}}{D}\,\frac{\sqrt{\dfrac{\varrho_2}{\varrho_1}} + \dfrac{D}{d}}{\log \dfrac{D}{d}}. \quad \dots\dots\dots \quad (5)$$

The second fraction in this expression depends on the ratio D/d and for a certain value of this ratio it passes through a (fairly flat) minimum. This optimum ratio is $D/d = 3.6$ for $\varrho_2/\varrho_1 = 1$, the usual case (inner and outer conductors made of same material). Thus, if we regard D as fixed, the cable attenuation will attain a minimum if d is made equal to $D/3.6$, in accordance with the condition (1). When the outer conductor is of aluminium and the inner of copper, $\varrho_2/\varrho_1 = 1.64$ and the condition of minimum attenuation occurs when $D/d = 3.8$.

Substituting for the second fraction in (5) the minimum value $4.6/\log_e 3.6$ and, taking into account the proportionality factors, not embodied for simplicity's sake in the above formulae, we finally obtain equation (3) for the attenuation. (In this case it is assumed that both conductors are of copper, for which $\varrho = 0.01745 \times 10^{-6}$ ohm metres, and that the dielectric is air ($\varepsilon_r = 1$).) By choosing D/d, likewise the inductance L and, when the relative dielectric constant ε_r of the insulation in the cable is known, the capacitance C and the characteristic impedance $\sqrt{L/C}$ of the coaxial tube are fixed. The characteristic impedance for a cable of optimum design with air dielectric is 77 ohms. With a typical construction using spacer beads or the equivalent, this figure is reduced to 75 ohms and with solid polythene insulation to 52 ohms.

The leakance G, which is caused by dielectric losses in the insulating medium, may be expressed in the form of a loss angle δ for the dielectric, by the equation:

$$G = \omega C \tan \delta \,. \qquad (6)$$

With most insulating materials used for coaxial cables, $\tan \delta$ does not depend on frequency within the frequency range of interest.

Since C, except for being proportional to ε_r, only depends on the ratio D/d, which has already been chosen for optimum resistance attenuation, it follows that, with this optimum design, G will be proportional to $f \cdot \varepsilon_r \cdot \tan \delta$. For polythene ε_r is not large (about 2-3) and dielectric losses are extremely small: a value $\tan \delta \approx 3 \times 10^{-4}$ is usually assumed. This leads to a leakance attenuation component which may be disregarded in ordinary cable construction, notwithstanding the fact that this component increases with frequency at a greater rate than the resistance attenuation.

Summary. Development of carrier telephone systems has been directed towards increasing the number of channels per physical circuit up to the economic limit in order to reduce the line cost per channel. In "quad" cables, 48 channels per pair is now normal practice (maximum frequency transmitted 204 kc/s). It is hoped to extend this figure in the future to 60, 72 or even 120 channels per pair. Further progress in quad cable development is limited by crosstalk which becomes more serious at the higher frequencies. On this score, things are fundamentally better with coaxial cable; owing to the screening effect of the outer conductor, crosstalk in this case decreases with increasing frequency. Crosstalk at frequencies below 60 kc/s is serious on coaxial cables and consequently this is normally the lowest frequency transmitted, but there is in this respect no upper frequency limit. Coaxial systems at present in operation in Great Britain and the U.S.A. transmit as many as 600-660 channels per pair, and standardisation by the C.C.I.F. provides for a future increase to 960 (maximum frequency transmitted 4 Mc/s). An economic limit is imposed by level considerations which lead to a reduction in the repeater spacing not commensurate with the number of extra channels obtained. This economic limit has not yet been reached, the number of channels being actually restricted due to the aggravation with increasing frequency of requirements in respect of reliability, stability and linearity of the repeaters. Finally, the problem of interconnection is discussed, this being inherent in all carrier systems having few physical circuits each carrying a large number of channels.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS OF THE N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk * can be obtained free of charge upon application to the address on the back cover.

1995: J. J. Went: The value of the spontaneous magnetization of binary nickel alloys as a function of temperature (Physica **17**, 596-602, 1951, No. 6).

The J_s versus T curve for pure nickel is more concave towards the T-axis than that for any binary nickel alloy, with only one exception, viz. completely ordered alloys, such as slowly-cooled Ni_3Fe. For pure nickel the form of the J_s versus T curve can be explained by the occurrence of an order-disorder phenomenon of the magnetic moments, where the latter can be placed only parallel or anti-parallel. It is suggested that all the other J_s versus T curves must be explained as being a result of this order-disorder phenomenon and of the statistical fluctuations of the concentration of dissolved atoms. In the case of Ni_3Fe two different order-disorder phenomena (a crystallographic and a magnetic one) act simultaneously.

1996: J. Smit: Magnetoresistance of ferromagnetic metals and alloys at low temperatures (Physica **17**, 612-627, 1951, No. 6).

The magnetoresistance of pure Ni and Fe, of Ni-Fe-, Ni-Co-, and Ni-Cu-alloys and of Heusler's alloy has been measured at room temperature and at temperatures of liquid nitrogen and liquid hydrogen. The behaviour of the pure metals is essentially different from that of the alloys. A maximum in the magnetoresistance is observed at low temperatures for alloys having about one Bohr magneton per atom. The positive difference between the longitudinal and the transversal resistance can be explained by means of the spin-orbit interaction.

At low temperatures the pure metals show an increase in resistance with increasing field just as the non-ferromagnetic metals. From this the value of the internal field, acting on the conduction electrons, could be determined, and was found to be approximately equal to the flux density B.

1997: J. A. Haringx: De instabiliteit van inwendig op druk belaste dunwandige cylinders (De Ingenieur **63**, O 39-O 41, 1951, No. 29). (The instability of thin-walled cylinders subjected to internal pressure; in Dutch.)

It is shown that under certain conditions a thin-walled cylinder may buckle when subjected to internal pressure. The critical value of this pressure can easily be calculated on account of Euler's well-known formula. Most of the formulae for pressure-loaded cylinders given in current textbooks, however, fail to predict this behaviour and should therefore be applied with caution for great lengths and/or for high pressures.

1998: J. A. Haringx: De instabiliteit van inwendig op druk belaste ronde balgen (De Ingenieur **63**, O 42-O 44, 1951, No. 29). (The instability of cylindrical bellows subjected to internal pressures; in Dutch.)

Like thin-walled cylinders, dealt with in a previous paper (see No. **1997**), also bellows may become unstable when loaded by internal pressure. The critical value of this pressure, which is governed, through Euler's well-known formula, by the rigidity of the bellows with respect to bending, is computed only for rectangularly shaped corrugations. The result found has been checked experimentally.

1999: C. G. Koops: On the dispersion of resistivity and dielectric constant of some semiconductors at audiofrequencies (Phys. Rev. **83**, 121-124, 1951, No. 1).

Semi-conducting $Ni_{0,4}Zn_{0,6}Fe_2O_4$, prepared in different ways, has been investigated. It appeared that the A.C. resistivity and the apparent dielectric constant of the material shows a dispersion, which can be explained satisfactorily with the help of a simple model of the solid: there should be well-conducting grains, separated by layers of lower conductivity. Dispersion formulas are given. There is good agreement between experiment and theory.

2000: E. J. W. Verwey: Oxidic semi-conductors (from: Semi-conducting materials, Butterworth's Scientific Publications Ltd., London 1951, pp. 151-161).

Non-conducting oxidic materials may be rendered conductive either by preparing solid solutions with oxides of high conductivity (e.g. a poorly conducting spinel with Fe_3O_4) or by introducing ions of deviating valency into the lattice (method of controlled valency). A comparison is made between semi-conductors obtained is this way and elemental semi-conductors such as silicon and germanium. The conductive properties of polycrystalline semi-conductors as a function of frequency are discussed, especially the influence of non-conducting thin layers at the grain boundaries (see these abstracts, Nos. **1738, 1739, 1845, 1914,** and Philips techn. Rev. **9**, 36-54, 239-248, 1947; **13**, 90-95, 1951, No. 2).

2001: J. Volger: Some properties of mixed lanthanum and strontium manganites (from: Semiconducting materials, Butterworth's Scientific Publications Ltd., London 1951, pp. 162-171).

Discussion of electrical properties of manganites $XMnO_3$. This material is an example of a polycrystalline semi-conductor prepared according to the principle of controlled valency (replacing the trivalent ion X, e.g. La, by a bivalent ion, such as Sr). There exists a close analogy between the resistivity and the ferromagnetic properties. Discussion of anomalies at the Curie temperature, of the influence of frequency, electric field and external magnetic field on the resistivity, of the thermo-electric properties and the Hall effect.

2002: H. Bremmer: On the diffraction theory of Gaussian optics (Comm. pure and appl. Math. 4, 61-74, 1951, No. 1).

The diffraction theory of optical imaging is developed for objects with arbitrary structure. The theory is based on rigorous solutions of the wave equation instead of the conventional approximation of Kirchoff's formula. The similarity of the wave functions in the object plane and in the corresponding paraxial image plane (Gaussian systems with unlimited aperture) proves to be connected with Neumann's integral theorem for Bessel functions (instead of the Fourier identity as in Kirchhoff's approximations). Another solution accounts for the effects of optical aberrations and of limited apertures.

2003: H. Bremmer: The W. K. B. approximation as the first term of a geometric-optical series (Comm. pure and appl. Math. 4, 104-115, 1951, No. 1).

The W.K.B. approximation of the solution of $y'' + k^2(x)\, y = 0$ is derived from a discontinuous model of an inhomogeneous medium. Higher approximations are found by considering multiple reflections. The solutions of different order form a series, the convergence of which is discussed. The well-known insufficiency of the W.K.B. approximation in the neighbourhood of zero's of $k(x)$ can be interpreted as a very slow convergence of the series mentioned.

2004: W. Ch. van Geel: On rectifiers (Physica **17**, 761-776, 1951, No. 8).

Experiments are described in which rectifiers were obtained from combinations of metals, semiconductors and intermediate layers. The following combinations have been investigated: (*a*) the contact between an excess semiconductor and a deficit semiconductor; (*b*) the combination aluminium/aluminium oxide/semiconductor; and (*c*) the combination metal/resin layer/semiconductor. In all these cases rectification occurs. The suggestion

is put forward that, in all three cases, the contact between two layers with charge carriers of opposite sign is the cause of rectification.

R 173: J. C. Francken and R. Dorrestein: Paraxial image formation in the "magnetic" image iconoscope (Philips Res. Rep. **6**, 323-346, 1951, No. 5).

In this paper a method is described for computing paraxial rays in a cathode lens placed in a magnetic field. In order to judge the approximations made, the well-known derivation of the ray equation is given. The solutions of this equation are discussed and a physical interpretation is given. A special case, approximating the electron-optical system in the image iconoscope, is computed numerically. The electron trajectories thus found lead to a discussion of the imaging mechanism in these electron-optical systems. In some respects the mechanism appears to differ considerably from that in ordinary magnetic lenses.

R 174: G. J. Fortuin: Visual power and visibility, II, (Philips Res. Rep. **6**, 347-371, 1951, No. 5).

See **R 170**. This part deals with the distribution of the visual power and of the visual types in age groups. Further the physiological meaning of the constant factors in the quantitative definition of visual power is discussed. One of these factors represents the lowest field brightness that allows perception of a dark object, another represents the brightness at which rod vision changes into cone vision and vice versa.

Visibility is the subject of the last chapter. Three possible definitions were tested by some standards, but only the ratio between the actual size of the object and the size of the smallest object perceptible under equal conditions (the so-called size-reduction factor) complied with our demands.

R 175: J. L. H. Jonker: The angular distribution of the secondary electrons of nickel (Philips Res. Rep. **6**, 372-387, 1951, No. 5).

The common equipment for measuring secondary emission (disc-shaped or spherical electrode within a sphere) is not suitable to obtain data about the angular distribution of the secondary electrons. To this aim an electrode system with two concentric spheres was constructed in order to obtain a really radial retarding electrostatic field, with which the behaviour of the secondary electrons with different velocities could be studied. The distribution of the secondary electrons (slow genuine secondary electrons, secondary electrons with moderate velocities, and rapid reflected electrons) was measured as a function of the angle of incidence and of the bombardment voltage of the primary electrons. The construction of the measuring tube, the method of measuring and the results obtained are discussed.

R 176: N. Warmoltz: The time lag of an ignitron (Philips Res. Rep. **6**, 388-400, 1951, No. 5).

To ignite periodically a discharge in a rectifier with a mercury-pool cathode, a current is sent through a semiconductive rod partly immersed in the mercury. The time lag in starting the discharge is measured on an oscillograph when a charged capacitor is switched by a thyratron onto the igniter. This is done for igniters of widely varying resistance in a liquid and a solid mercury cathode and in a liquid and a solid tin cathode. Also the influence of the gas pressure in the tube is investigated. The results are compared with those of Slepian and Ludwig and those of Dow and Powers. Finally the field emission theory of Slepian and Ludwig and the thermal theory of Mierdel are discussed, whereby it turns out that the measurements of the time lags with liquid and solid cathodes fit best in the thermal theory.

ERRATUM

The colour reproduction of a painting by Giovanni Bellini on page 77 in the preceding issue of this Review is by courtesy of the Trustees of the National Gallery, London. The editors regret that owing to a mistake this statement was omitted in the subscript.